

## **CHAPTER 6**

### *Power Control with Thyristors and Triacs*

- 6.1 Using Thyristors and Triacs*
- 6.2 Thyristor and Triac Applications*
- 6.3 Hi-Com Triacs*



## Using Thyristors and Triacs





### 6.1.1 Introduction to Thyristors and Triacs

#### Brief summary of the thyristor family

The term thyristor is a generic name for a semiconductor switch having four or more layers and is, in essence, a p-n-p-n sandwich. Thyristors form a large family and it is helpful to consider the constituents which determine the type of any given thyristor. If an ohmic connection is made to the first p region and the last n region, and no other connection is made, the device is a diode thyristor. If an additional ohmic connection is made to the intermediate n region (n gate type) or the intermediate p region (p gate type), the device is a triode thyristor. If an ohmic connection is made to both intermediate regions, the device is a tetrode thyristor. All such devices have a forward characteristic of the general form shown in Fig. 1.

There are three types of thyristor reverse characteristic: blocking (as in normal diodes), conducting (large reverse currents at low reverse voltages) and approximate mirror image of the forward characteristic (bidirectional thyristors). Reverse blocking devices usually have four layers or less whereas reverse conducting and mirror image devices usually have five layers.

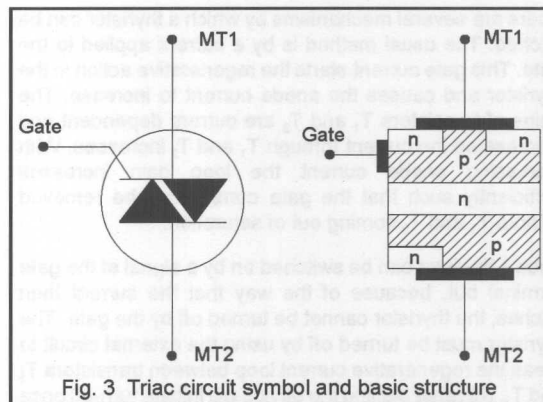
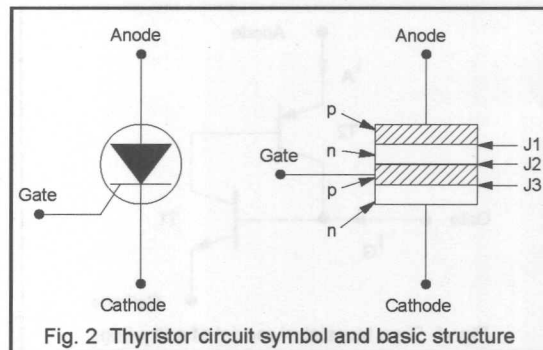
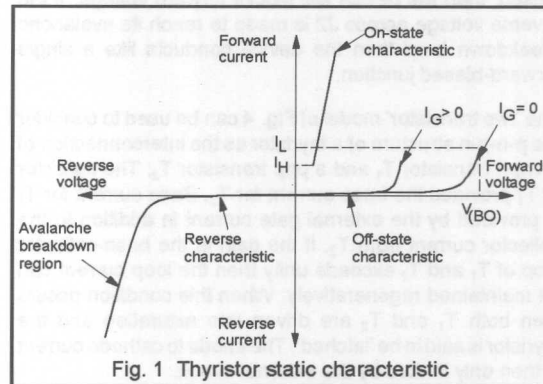
The simplest thyristor structure, and the most common, is the reverse blocking triode thyristor (usually simply referred to as the 'thyristor' or SCR 'silicon controlled rectifier'). Its circuit symbol and basic structure are shown in Fig. 2.

The most complex common thyristor structure is the bidirectional triode thyristor, or triac. The triac (shown in Fig. 3) is able to pass current bidirectionally and is therefore an a.c. power control device. Its performance is that of a pair of thyristors in anti-parallel with a single gate terminal. The triac needs only one heatsink, but this must be large enough to remove the heat caused by bidirectional current flow. Triac gate triggering circuits must be designed with care to ensure that unwanted conduction, i.e. loss of control, does not occur when triggering lasts too long.

Thyristors and triacs are both bipolar devices. They have very low on-state voltages but, because the minority charge carriers in the devices must be removed before they can block an applied voltage, the switching times are comparatively long. This limits thyristor switching circuits to low frequency applications. Triacs are used almost exclusively at mains supply frequencies of 50 or 60Hz, while in some applications this extends up to the 400Hz supply frequency as used in aircraft.

The voltage blocking capabilities of thyristors and triacs are quite high: the highest voltage rating for the Philips range is 800V, while the currents ( $I_{T(RMS)}$ ) range from 0.8A to 25A.

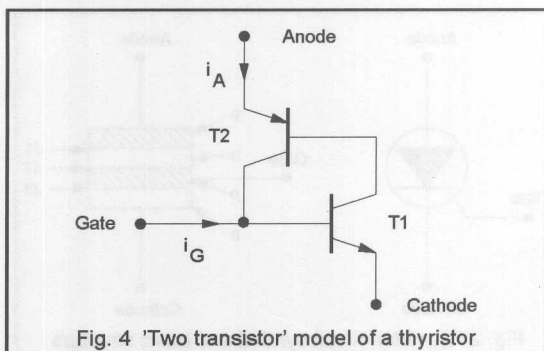
The devices are available as surface mount components, or as non-isolated or isolated discrete devices, depending on the device rating.



## Thyristor operation

The operation of the thyristor can be understood from Fig. 2. When the thyristor cathode is more positive than the anode then junctions J1 and J3 are reverse biased and the device blocks. When the anode is more positive than the cathode, junctions J1 and J3 are forward biased. As J2 is reverse biased, then the device still blocks forward voltage. If the reverse voltage across J2 is made to reach its avalanche breakdown level then the device conducts like a single forward-biased junction.

The 'two transistor' model of Fig. 4 can be used to consider the p-n-p-n structure of a thyristor as the interconnection of an npn transistor  $T_1$  and a pnp transistor  $T_2$ . The collector of  $T_1$  provides the base current for  $T_2$ . Base current for  $T_1$  is provided by the external gate current in addition to the collector current from  $T_2$ . If the gain in the base-collector loop of  $T_1$  and  $T_2$  exceeds unity then the loop current can be maintained regeneratively. When this condition occurs then both  $T_1$  and  $T_2$  are driven into saturation and the thyristor is said to be 'latched'. The anode to cathode current is then only limited by the external circuit.



There are several mechanisms by which a thyristor can be latched. The usual method is by a current applied to the gate. This gate current starts the regenerative action in the thyristor and causes the anode current to increase. The gains of transistors  $T_1$  and  $T_2$  are current dependent and increase as the current through  $T_1$  and  $T_2$  increases. With increasing anode current the loop gain increases sufficiently such that the gate current can be removed without  $T_1$  and  $T_2$  coming out of saturation.

Thus a thyristor can be switched on by a signal at the gate terminal but, because of the way that the current then latches, the thyristor cannot be turned off by the gate. The thyristor must be turned off by using the external circuit to break the regenerative current loop between transistors  $T_1$  and  $T_2$ . Reverse biasing the device will initiate turn-off once the anode current drops below a minimum specified value, called the holding current value,  $I_H$ .

## Thyristor turn-on methods

### Turn-on by exceeding the breakover voltage

When the breakover voltage,  $V_{BO}$ , across a thyristor is exceeded, the thyristor turns on. The breakover voltage of a thyristor will be greater than the rated maximum voltage of the device. At the breakover voltage the value of the thyristor anode current is called the latching current,  $I_L$ .

Breakover voltage triggering is not normally used as a triggering method, and most circuit designs attempt to avoid its occurrence. When a thyristor is triggered by exceeding  $V_{BO}$  the fall time of the forward voltage is quite low (about 1/20th of the time taken when the thyristor is gate-triggered). As a general rule, however, although a thyristor switches faster with  $V_{BO}$  turn-on than with gate turn-on, the permitted  $di/dt$  for breakover voltage turn-on is lower.

### Turn-on by leakage current

As the junction temperature of a thyristor rises, the leakage current also increases. Eventually, if the junction temperature is allowed to rise sufficiently, leakage current would become large enough to initiate latching of the regenerative loop of the thyristor and allow forward conduction. At a certain critical temperature (above  $T_{J(max)}$ ) the thyristor will not support any blocking voltage at all.

### Turn-on by $dV/dt$

Any p-n junction has capacitance - the larger the junction area the larger the capacitance. If a voltage ramp is applied across the anode-to-cathode of a p-n-p-n device, a current will flow in the device to charge the device capacitance according to the relation:

$$i_c = C \cdot \frac{dv}{dt} \quad (1)$$

If the charging current becomes large enough, the density of moving current carriers in the device induces switch-on.

### Turn-on by gate triggering

Gate triggering is the usual method of turning a thyristor on. Application of current to the thyristor gate initiates the latching mechanism discussed in the previous section. The characteristic of Fig. 1 showed that the thyristor will switch to its on-state condition with forward bias voltages less than  $V_{BO}$  when the gate current is greater than zero. The gate current and voltage requirements which ensure triggering of a particular device are always quoted in the device data. As thyristor triggering characteristics are temperature dependant, the amplitude and duration of the gate pulse must be sufficient to ensure that the thyristor latches under all possible conditions.

During gate turn-on, the rate of rise of thyristor anode current  $di_F/dt$  is determined by the external circuit conditions. However, the whole active area of the thyristor (or triac) cannot be turned on simultaneously: the area nearest to the gate turns on first, followed by the remainder of the device. At turn-on it is important that the rate of rise of current does not exceed the specified rating. If  $di_F/dt$  is excessive then only a limited area of the device will have been turned on as the anode current increases. The resulting localised heating of the device will cause degradation and could lead to eventual device failure.

A suitably high gate current and large rate of rise of gate current ( $di_G/dt$ ) ensures that the thyristor turns on quickly (providing that the gate power ratings are not exceeded) thus increasing the thyristor turn-on  $di/dt$  capability. Once the thyristor has latched then the gate drive can be reduced or removed completely. Gate power dissipation can also be reduced by triggering the thyristor using a pulsed signal.

### Triac operation

The triac can be considered as two thyristors connected in antiparallel as shown in Fig. 5. The single gate terminal is common to both thyristors. The main terminals MT1 and MT2 are connected to both p and n regions of the device and the current path through the layers of the device depends upon the polarity of the applied voltage between the main terminals. The device polarity is usually described with reference to MT1, where the term MT2+ denotes that terminal MT2 is positive with respect to terminal MT1.

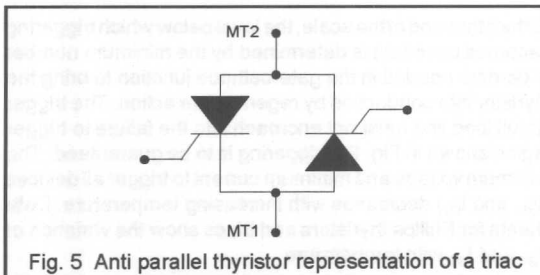


Fig. 5 Anti parallel thyristor representation of a triac

The on-state characteristic of the triac is similar to that of a thyristor and is shown in Fig. 6. Table 1 and Fig. 7 summarise the different gate triggering configurations for triacs.

Due to the physical layout of the semiconductor layers in a triac, the values of latching current ( $I_L$ ), holding current ( $I_H$ ) and gate trigger current ( $I_{GT}$ ) vary slightly between the different operating quadrants. In general, for any triac, the latching current is slightly higher in the second (MT2+, G-) quadrant than the other quadrants, whilst the gate trigger current is slightly higher in fourth (MT2-, G+) quadrant.

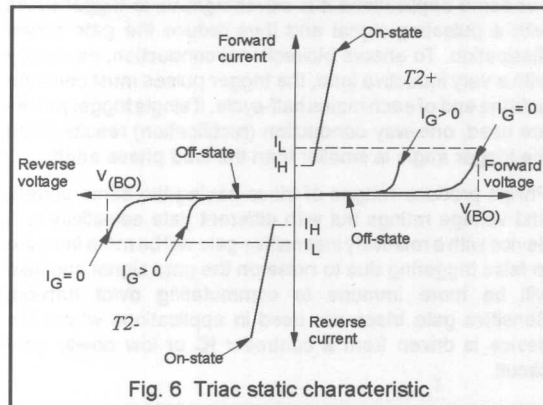


Fig. 6 Triac static characteristic

Quadrant	Polarity of MT2 wrt MT1	Gate polarity
1 (1+)	MT2+	G+
2 (1-)	MT2+	G-
3 (3-)	MT2-	G-
4 (3+)	MT2-	G+

Table 1. Operating quadrants for triacs

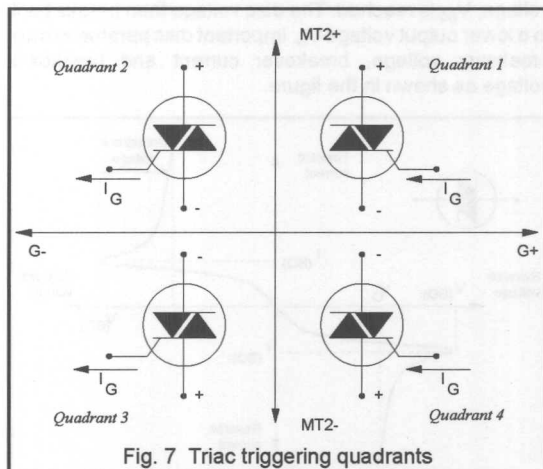


Fig. 7 Triac triggering quadrants

For applications where the gate sensitivity is critical and where the device must trigger reliably and evenly for applied voltages in both directions it may be preferable to use a negative current triggering circuit. If the gate drive circuit is arranged so that only quadrants 2 and 3 are used (i.e. G-operation) then the triac is never used in the fourth quadrant where  $I_{GT}$  is highest.



For some applications it is advantageous to trigger triacs with a pulsating signal and thus reduce the gate power dissipation. To ensure bidirectional conduction, especially with a very inductive load, the trigger pulses must continue until the end of each mains half-cycle. If single trigger pulses are used, one-way conduction (rectification) results when the trigger angle is smaller than the load phase angle.

Philips produce ranges of triacs having the same current and voltage ratings but with different gate sensitivities. A device with a relatively insensitive gate will be more immune to false triggering due to noise on the gate signal and also will be more immune to commutating  $dv/dt$  turn-on. Sensitive gate triacs are used in applications where the device is driven from a controller IC or low power gate circuit.

### The diac

It is also worthwhile to consider the operation and characteristics of the diac in the context of multilayer bipolar devices. The diac is more strictly a transistor than a thyristor, but has an important role in many thyristor and triac triggering circuits. It is manufactured by diffusing an n-type impurity into both sides of a p-type slice to give a two terminal device with symmetrical electrical characteristics. As shown in the characteristic of Fig. 8, the diac blocks applied voltages in either direction until the breakover voltage,  $V_{BO}$ , is reached. The diac voltage then breaks back to a lower output voltage  $V_O$ . Important diac parameters are breakover voltage, breakover current and breakback voltage as shown in the figure.

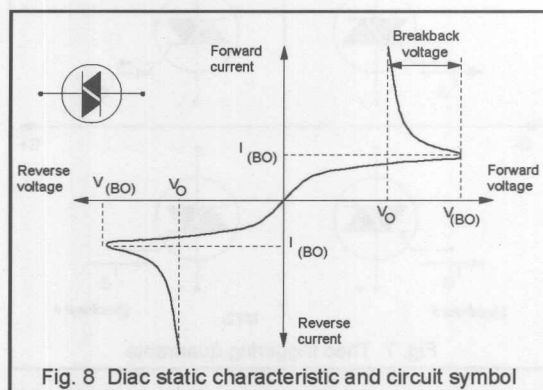


Fig. 8 Diac static characteristic and circuit symbol

### Gate requirements for triggering

To a first approximation, the gate-to-cathode junction of a thyristor or triac acts as a p-n diode. The forward characteristic is as shown in Fig. 9. For a given thyristor type there will be a spread in forward characteristics of gate junctions and a spread with temperature.

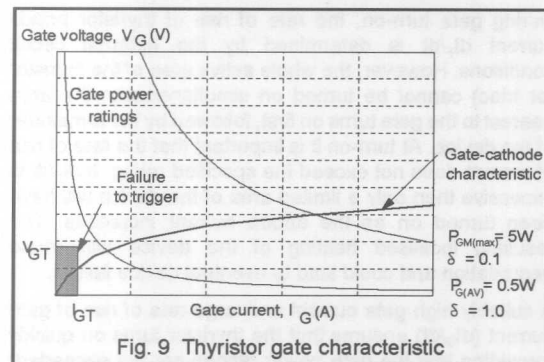


Fig. 9 Thyristor gate characteristic

The gate triggering characteristic is limited by the gate power dissipation. Figure 9 also shows the continuous power rating curve ( $P_{G(AV)}=0.5W$ ) for a typical device and the peak gate power curve ( $P_{GM(max)}=5W$ ). When designing a gate circuit to reliably trigger a triac or thyristor the gate signal must lie on a locus within the area of certain device triggering. Continuous steady operation would demand that the 0.5W curve be used to limit the load line of the gate drive circuit. For pulsed operation the triggering locus can be increased. If the 5W peak gate power curve is used, the duty cycle must not exceed

$$\delta_{max} = \frac{P_{G(AV)}}{P_{GM}} = \frac{0.5}{5} = 0.1 \quad (2)$$

At the other end of the scale, the level below which triggering becomes uncertain is determined by the minimum number of carriers needed in the gate-cathode junction to bring the thyristor into conduction by regenerative action. The trigger circuit load line must not encroach into the failure to trigger region shown in Fig. 9 if triggering is to be guaranteed. The minimum voltage and minimum current to trigger all devices ( $V_{GT}$  and  $I_{GT}$ ) decreases with increasing temperature. Data sheets for Philips thyristors and triacs show the variation of  $V_{GT}$  and  $I_{GT}$  with temperature.

### Thyristor commutation

A thyristor turns off by a mechanism known as 'natural turn-off', that is, when the main anode-cathode current drops below the holding value. It is important to remember, however, that the thyristor will turn on again if the reapplied forward voltage occurs before a minimum time period has elapsed; this is because the charge carriers in the thyristor at the time of turn-off take a finite time to recombine. Thyristor turn-off is achieved by two main methods - self commutation or external commutation.

**Self Commutation**

In self-commutation circuits the thyristor will automatically turn off at a predetermined time after triggering. The thyristor conduction period is determined by a property of the commutation circuit, such as the resonant cycle of an LC-circuit or the Volt-Second capability of a saturable inductor. The energy needed for commutation is delivered by a capacitor included in the commutation circuit.

**LC circuit in series with the thyristor**

When the thyristor is triggered, the resulting main current excites the resonant circuit. After half a resonant cycle, the LC circuit starts to reverse the anode current and turns the thyristor off. The thyristor conduction interval is half a resonant cycle. It is essential for proper commutation that the resonant circuit be less than critically damped. Fig. 10 shows the circuit diagram and the relevant waveforms for this arrangement.

**LC Circuit in parallel with the thyristor**

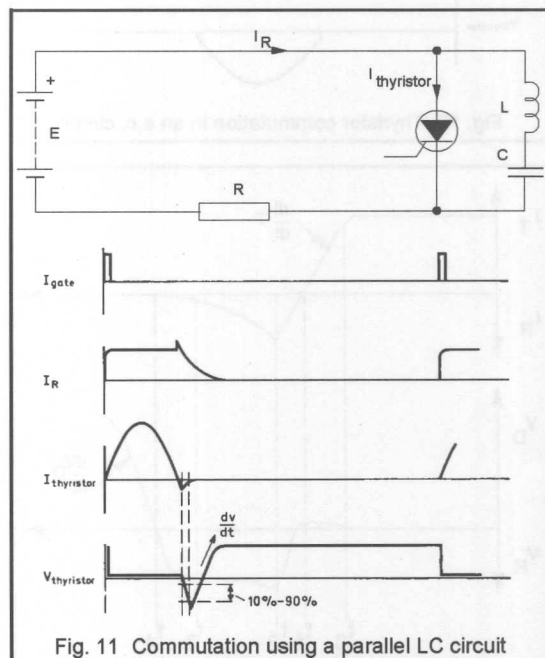
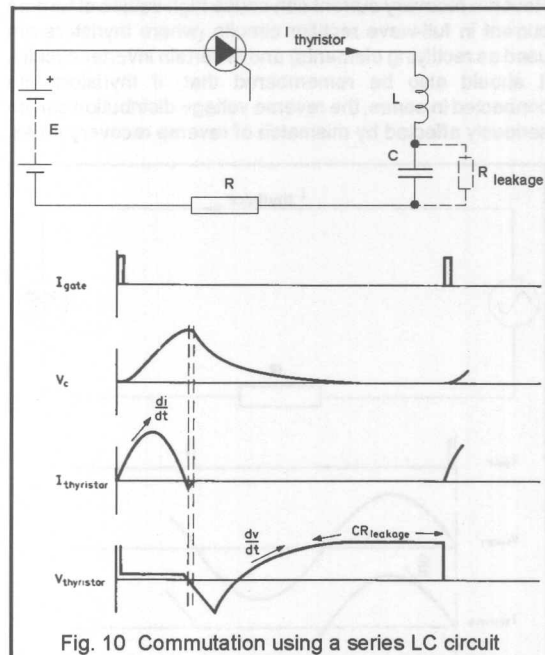
Initially the capacitor charges to the supply voltage. When the thyristor is triggered the load current flows but at the same time the capacitor discharges through the thyristor in the forward direction. When the capacitor has discharged (i.e. after one resonant half-cycle of the LC circuit), it begins to charge in the opposite direction and, when this charging current is greater than the thyristor forward current, the thyristor turns off. The circuit diagram and commutation waveforms are shown in Fig. 11.

**External commutation**

If the supply is an alternating voltage, the thyristor can conduct only during the positive half cycle. The thyristor naturally switches off at the end of each positive half cycle. The circuit and device waveforms for this method of commutation are shown in Fig. 12. It is important to ensure that the duration of a half cycle is greater than the thyristor turn-off time.

**Reverse recovery**

In typical thyristors the reverse recovery time is of the order of a few micro-seconds. This time increases with increase of forward current and also increases as the forward current decay rate,  $di/dt$ , decreases. Reverse recovery time is the period during which reverse recovery current flows ( $t_1$  to  $t_3$  in Fig. 13) and it is the period between the point at which forward current ceases and the earliest point at which the reverse recovery current has dropped to 10% of its peak value.



Reverse recovery current can cause high values of turn-on current in full-wave rectifier circuits (where thyristors are used as rectifying elements) and in certain inverter circuits. It should also be remembered that, if thyristors are connected in series, the reverse voltage distribution can be seriously affected by mismatch of reverse recovery times.

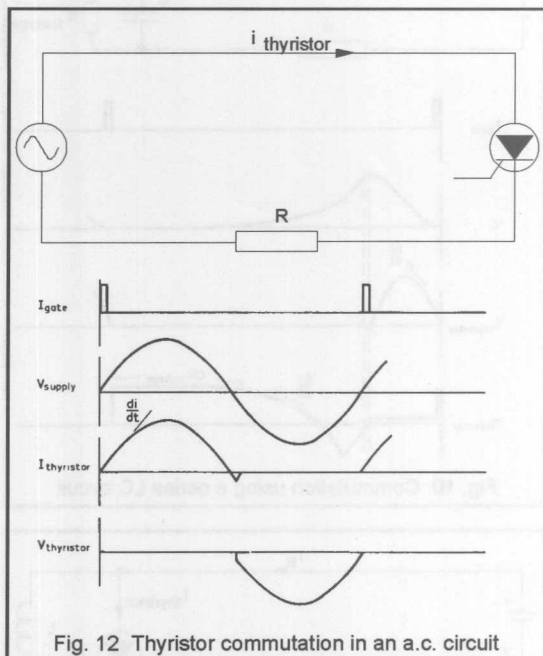


Fig. 12 Thyristor commutation in an a.c. circuit

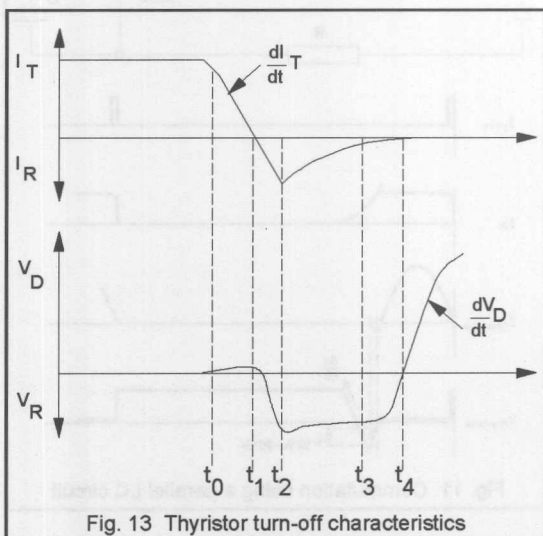


Fig. 13 Thyristor turn-off characteristics

### Turn-off time

Turn-off time is the interval between the instant when thyristor current reverses and the point at which the thyristor can block reapplied forward voltage ( $t_1$  to  $t_4$  in Fig. 13). If forward voltage is applied to a thyristor too soon after the main current has ceased to flow, the thyristor will turn on. The circuit commutated turn-off time increases with:

- junction temperature
- forward current amplitude
- rate of fall of forward current
- rate of rise of forward blocking voltage
- forward blocking voltage.

Thus the turn-off time is specified for defined operating conditions. Circuit turn-off time is the turn-off time that the circuit presents to the thyristor; it must, of course, be greater than the thyristor turn-off time.

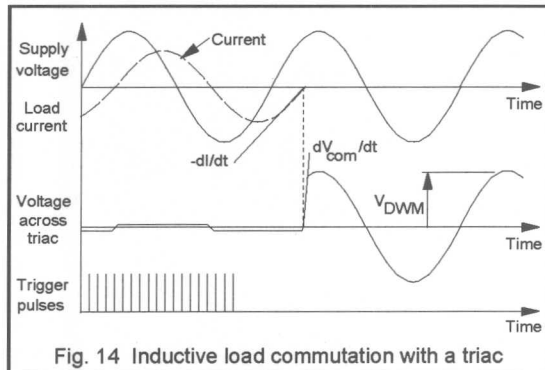
### Triac commutation

Unlike the thyristor, the triac can conduct irrespective of the polarity of the applied voltage. Thus the triac does not experience a circuit-imposed turn-off time which allows each anti-parallel thyristor to fully recover from its conducting state as it is reverse biased. As the voltage across the triac passes through zero and starts to increase, then the alternate thyristor of the triac can fail to block the applied voltage and immediately conduct in the opposite direction. Triac-controlled circuits therefore require careful design in order to ensure that the triac does not fail to commutate (switch off) at the end of each half-cycle as expected.

It is important to consider the commutation performance of devices in circuits where either  $di/dt$  or  $dV/dt$  can be large. In resistive load applications (e.g. lamp loads) current surges at turn-on or during temporary over-current conditions may introduce abnormally high rates of change of current which may cause the triac to fail to commutate. In inductive circuits, such as motor control applications or circuits where a dc load is controlled by a triac via a bridge rectifier, it is usually necessary to protect the triac against unwanted commutation due to  $dv_{(com)}/dt$ .

The commutating  $dv_{(com)}/dt$  limit for a triac is less than the static  $dv/dt$  limit because at commutation the recently conducting portion of the triac which is being switched off has introduced stored charge to the triac. The amount of stored charge depends upon the reverse recovery characteristics of the triac. It is significantly affected by junction temperature and the rate of fall of anode current prior to commutation ( $di_{(com)}/dt$ ). Following high rates of change of current the capacity of the triac to withstand high reapplied rates of change of voltage is reduced. Data sheet specifications for triacs give characteristics showing the

maximum allowable rate of rise of commutating voltage against device temperature and rate of fall of anode current which will not cause a device to trigger.



Consider the situation when a triac is conducting in one direction and the applied ac voltage changes polarity. For the case of an inductive load the current in the triac does not fall to its holding current level until some time later. This is shown in Fig. 14. At the time that the triac current has reached the holding current the mains voltage has risen to some value and so the triac must immediately block that voltage. The rate of rise of blocking voltage following commutation ( $dv_{com}/dt$ ) can be quite high.

The usual method is to place a  $dv/dt$ -limiting R-C snubber in parallel with the triac. Additionally, because commutating  $dv/dt$  turn-on is dependent upon the rate of fall of triac current, then in circuits with large rates of change of anode current, the ability of a triac to withstand high rates of rise of reapplied voltage is improved by limiting the  $di/dt$  using a series inductor. This topic is discussed more fully in the section entitled 'Using thyristors and triacs'.

## Conclusions

This article has presented the basic parameters and characteristics of triacs and thyristors and shown how the structure of the devices determines their operation. Important turn-on and turn-off conditions and limitations of the devices have been presented in order to demonstrate the capabilities of the devices and show the designer those areas which require careful consideration. The device characteristics which determine gate triggering requirements of thyristors and triacs have been presented.

Subsequent articles in this chapter will deal with the use, operation and limitations of thyristors and triacs in practical applications, and will present some detailed design and operational considerations for thyristors and triacs in phase control and integral cycle control applications.





## 6.1.2 Using Thyristors and Triacs

This chapter is concerned with the uses, operation and protection of thyristors and triacs. Two types of circuit cover the vast majority of applications for thyristors and triacs: static switching circuits and phase control circuits. The characteristics and uses of these two types of circuit will be discussed. Various gate drive circuits and protection circuits for thyristor and triacs are also presented. The use of these circuits will enable designers to operate the devices reliably and within their specified limits.

### Thyristor and triac control techniques

There are two main techniques of controlling thyristors and triacs - on-off triggering (or static switching) and phase control. In on-off triggering, the power switch is allowed to conduct for a certain number of half-cycles and then it is kept off for a number of half-cycles. Thus, by varying the ratio of "on-time" to "off-time", the average power supplied to the load can be controlled. The switching device either completely activates or deactivates the load circuit. In phase control circuits, the thyristor or triac is triggered into conduction at some point after the start of each half-cycle. Control is achieved on a cycle-by-cycle basis by variation of the point in the cycle at which the thyristor is triggered.

### Static switching applications

Thyristors and triacs are the ideal power switching devices for many high power circuits such as heaters, enabling the load to be controlled by a low power signal, in place of a relay or other electro-mechanical switch.

In a high power circuit where the power switch may connect or disconnect the load at any point of the mains cycle then large amounts of RFI (radio frequency interference) are likely to occur at the instants of switching. The large variations in load may also cause disruptions to the supply voltage. The RFI and voltage variation produced by high power switching in a.c. mains circuits is unacceptable in many environments and is controlled by statutory limits. The limits depend upon the type of environment (industrial or domestic) and the rating of the load being switched.

RFI occurs at any time when there is a step change in current caused by the closing of a switch (mechanical or semiconductor). The energy levels of this interference can be quite high in circuits such as heating elements. However, if the switch is closed at the moment the supply voltage passes through zero there is no step rise in current and thus no radio frequency interference. Similarly, at turn-off, a large amount of high frequency interference can be caused by di/dt imposed voltage transients in inductive circuits.

Circuit-generated RFI can be almost completely eliminated by ensuring that the turn-on switching instants correspond to the zero-crossing points of the a.c. mains supply. This technique is known as synchronous (or zero voltage) switching control as opposed to the technique of allowing the switching points to occur at any time during the a.c. cycle, which is referred to as asynchronous control.

In a.c. circuits using thyristors and triacs the devices naturally switch off when the current falls below the device holding current. Thus turn-off RFI does not occur.

### Asynchronous control

In asynchronous control the thyristor or triac may be triggered at a point in the mains voltage other than the zero voltage crossover point. Asynchronous control circuits are usually relatively cheap but liable to produce RFI.

### Synchronous control

In synchronous control systems the switching instants are synchronised with zero crossings of the supply voltage. They also have the advantage that, as the thyristors conduct over complete half cycles, the power factor is very good. This method of power control is mostly used to control temperature. The repetition period,  $T$ , is adjusted to suit the controlled process (within statutory limits). Temperature ripple is eliminated when the repetition period is made much smaller than the thermal time constant of the system.

Figure 1 shows the principle of time-proportional control. RFI and turn-on di/dt are reduced, and the best power factor (sinusoidal load current) is obtained by triggering synchronously. The average power delivered to a resistive load,  $R_L$ , is proportional to  $t_{on}/T$  (i.e. linear control) and is given by equation 1.

$$P_{out} = \frac{V_{(RMS)}^2}{R_L} \cdot \frac{t_{on}}{T} \quad (1)$$

where:  $T$  is the controller repetition period  
 $t_{on}$  is controller 'on' time  
 $V_{(RMS)}$  is the rms a.c. input voltage.

Elsewhere in this handbook the operation of a controller i.c. (the TDA1023) is described. This device is specifically designed to implement time-proportional control of heaters using Philips triacs.

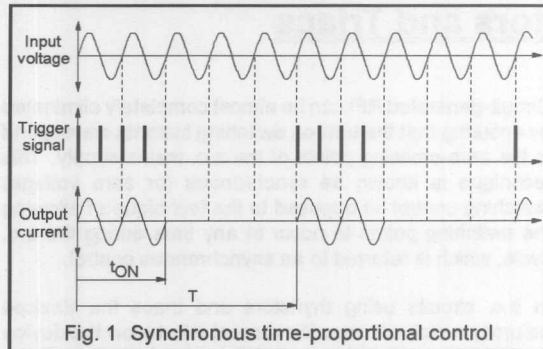


Fig. 1 Synchronous time-proportional control

### Phase control

Phase control circuits are used for low power applications such as lamp control or universal motor speed control, where RFI emissions can be filtered relatively easily. The power delivered to the load is controlled by the timing of the thyristor (or triac) turn-on point.

The two most common phase controller configurations are 'half wave control', where the controlling device is a single thyristor and 'full wave control', where the controlling device is a triac or a pair of anti-parallel thyristors. These two control strategies are considered in more detail below:

### Resistive loads

The operation of a phase controller with a resistive load is the simplest situation to analyse. Waveforms for a full wave controlled resistive load are shown in Fig. 2. The triac is triggered at angle  $\delta$ , and applies the supply voltage to the load. The triac then conducts for the remainder of the positive half-cycle, turning off when the anode current drops below the holding current, as the voltage becomes zero at  $\theta=180^\circ$ . The triac is then re-triggered at angle  $(180+\delta)^\circ$ , and conducts for the remainder of the negative half-cycle, turning off when its anode voltage becomes zero at  $360^\circ$ .

The sequence is repeated giving current pulses of alternating polarity which are fed to the load. The duration of each pulse is the conduction angle  $\alpha$ , that is  $(180-\delta)^\circ$ . The output power is therefore controlled by variation of the trigger angle  $\delta$ .

For all values of  $\alpha$  other than  $\alpha=180^\circ$  the load current is non-sinusoidal. Thus, because of the generation of harmonics, the power factor presented to the a.c. supply will be less than unity except when  $\delta=0$ .

For a sinusoidal current the rectified mean current,  $I_{T(AV)}$ , and the rms current,  $I_{T(RMS)}$ , are related to the peak current,  $I_{T(MAX)}$ , by equation 2.

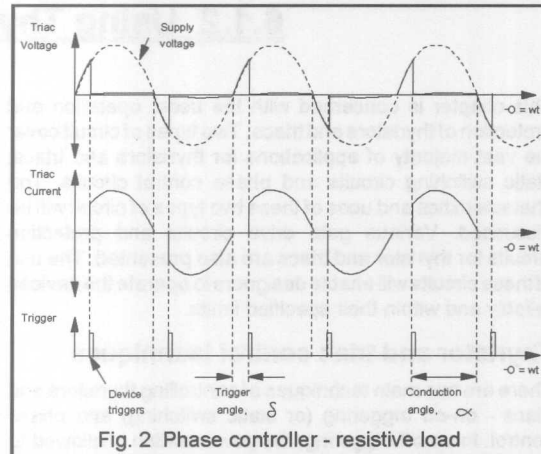


Fig. 2 Phase controller - resistive load

$$I_{T(AV)} = \frac{2I_{T(MAX)}}{\pi} = 0.637 I_{T(MAX)}$$

$$I_{T(RMS)} = \frac{I_{T(MAX)}}{\sqrt{2}} = 0.707 I_{T(MAX)} \quad (2)$$

where

$$I_{T(MAX)} = \frac{V_{T(MAX)}}{R_L} = \frac{\sqrt{2}V_{(RMS)}}{R_L} \quad (3)$$

From equation 2 the 'crest factor',  $c$ , (also known as the 'peak factor') of the current waveform is defined as:

$$\text{Crest factor, } c = \frac{I_{T(MAX)}}{I_{T(RMS)}} \quad (4)$$

The current 'form factor',  $a$ , is defined by:

$$\text{Form factor, } a = \frac{I_{T(RMS)}}{I_{T(AV)}} \quad (5)$$

Thus, for sinusoidal currents:

$$a = \frac{I_{T(RMS)}}{I_{T(AV)}} = 1.111; \quad c = \frac{I_{T(MAX)}}{I_{T(RMS)}} = 1.414 \quad (6)$$

For the non-sinusoidal waveforms which occur in a phase controlled circuit, the device currents are modified due to the delay which occurs before the power device is triggered. The crest factor of equation 4 and the form factor of equation 5 can be used to describe variation of the current waveshape from the sinusoidal case.

**Half wave controller**

Figure 3a) shows the simplest type of thyristor half-wave phase controller for a resistive load. The load current waveform is given in Fig. 3b). The variation of average load current,  $I_{T(AV)}$ , rms load current,  $I_{T(RMS)}$  and load power over the full period of the a.c. mains, with trigger angle are given in equation 7.

$$I_{T(AV)} = I_{T(AV)\max} \cdot \frac{(1 - \cos \alpha)}{2} \quad I_{T(AV)\max} = \frac{I_{T(MAX)}}{\pi}$$

$$I_{T(RMS)} = I_{T(RMS)\max} \left( \frac{\alpha - \frac{1}{2} \sin 2\alpha}{\pi} \right)^{\frac{1}{2}} \quad I_{T(RMS)\max} = \frac{I_{T(MAX)}}{2}$$

$$P_{(out)} = P_{(out)\max} \left( \frac{\alpha - \frac{1}{2} \sin 2\alpha}{\pi} \right) \quad P_{(out)\max} = \frac{I_{T(MAX)}^2 R_L}{4} \quad (7)$$

N.B. When using equation 7 all values of  $\alpha$  must be in radians. For each case the maximum value occurs when  $\alpha = 180^\circ$  ( $\alpha = \pi$  radians).

At  $\alpha = 180^\circ$  the crest factor and form factor for a half wave controller are given by:

$$a = \frac{I_{T(RMS)}}{I_{T(AV)}} = 1.571; \quad c = \frac{I_{T(MAX)}}{I_{T(RMS)}} = 2.0 \quad (8)$$

**Full wave controller**

Figure 4 shows the circuit and load current waveforms for a full-wave controller using two antiparallel thyristors, or a triac, as the controlling device. The variation of rectified mean current,  $I_{T(AV)}$ , rms current,  $I_{T(RMS)}$ , and load power with trigger angle are given by equation 9.

$$I_{T(AV)} = I_{T(AV)\max} \cdot \frac{(1 - \cos \alpha)}{2} \quad I_{T(AV)\max} = \frac{2I_{T(MAX)}}{\pi}$$

$$I_{T(RMS)} = I_{T(RMS)\max} \left( \frac{\alpha - \frac{1}{2} \sin 2\alpha}{\pi} \right)^{\frac{1}{2}} \quad I_{T(RMS)\max} = \frac{I_{T(MAX)}}{\sqrt{2}}$$

$$P_{(out)} = P_{(out)\max} \left( \frac{\alpha - \frac{1}{2} \sin 2\alpha}{\pi} \right) \quad P_{(out)\max} = \frac{I_{T(MAX)}^2 R_L}{2} \quad (9)$$

N.B. When using equation 9 all value of  $\alpha$  must be in radians. For each case the maximum value occurs when  $\alpha = 180^\circ$  ( $\alpha = \pi$  radians).

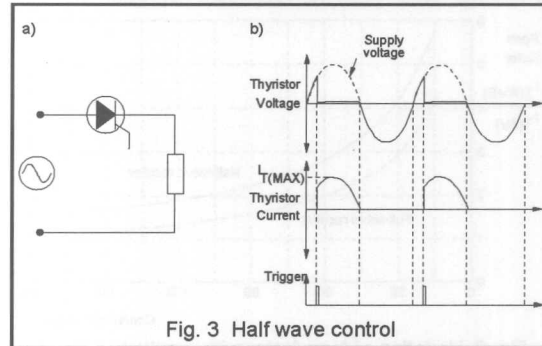


Fig. 3 Half wave control

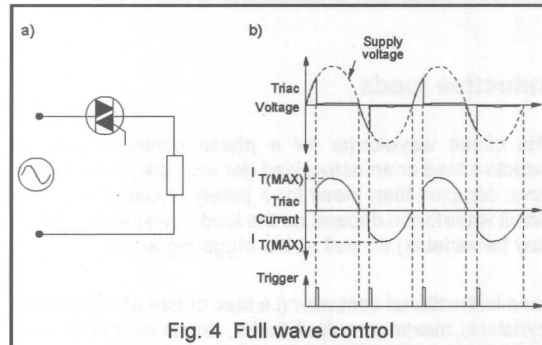


Fig. 4 Full wave control

The variation of normalised average current,  $I_{T(AV)}/I_{T(AV)\max}$ , rms current  $I_{T(RMS)}/I_{T(RMS)\max}$ , and power,  $P_{(out)}/P_{(out)\max}$ , for equations 7 and 9 are plotted in Fig. 5.

Figure 6 shows the variation of current form factor with conduction angle for the half wave controller and the full wave controller of Figs. 3 and 4.

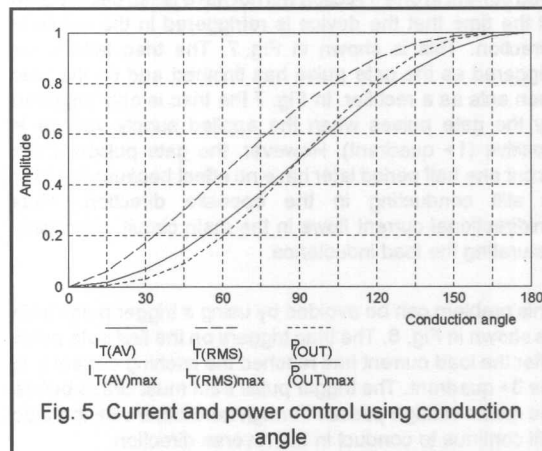


Fig. 5 Current and power control using conduction angle



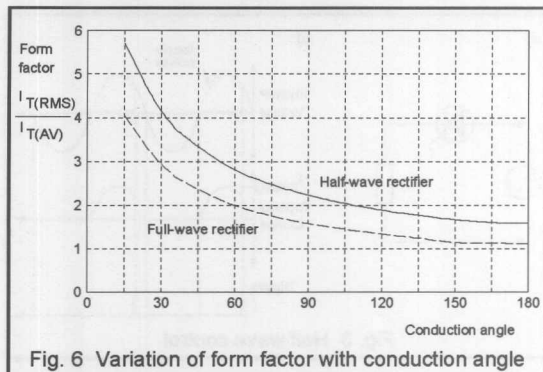


Fig. 6 Variation of form factor with conduction angle

### Inductive loads

The circuit waveforms for a phase controller with an inductive load or an active load (for example, a motor) are more complex than those for a purely resistive load. The circuit waveforms depend on the load power factor (which may be variable) as well as the triggering angle.

For a bidirectional controller (i.e. triac or pair of anti-parallel thyristors), maximum output, that is, sinusoidal load current, occurs when the trigger angle equals the phase angle. When the trigger angle,  $\delta$ , is greater than the load phase angle,  $\phi$ , then the load current will become discontinuous and the triac (or thyristor) will block some portion of the input voltage until it is retriggered.

If the trigger angle is less than the phase angle then the load current in one direction will not have fallen back to zero at the time that the device is retriggered in the opposite direction. This is shown in Fig. 7. The triac fails to be triggered as the gate pulse has finished and so the triac then acts as a rectifier. In Fig. 7 the triac is only triggered by the gate pulses when the applied supply voltage is positive (1+ quadrant). However, the gate pulses which occur one half period later have no effect because the triac is still conducting in the opposite direction. Thus unidirectional current flows in the main circuit, eventually saturating the load inductance.

This problem can be avoided by using a trigger pulse train as shown in Fig. 8. The triac triggers on the first gate pulse after the load current has reached the latching current  $I_L$  in the 3+ quadrant. The trigger pulse train must cease before the mains voltage passes through zero otherwise the triac will continue to conduct in the reverse direction.

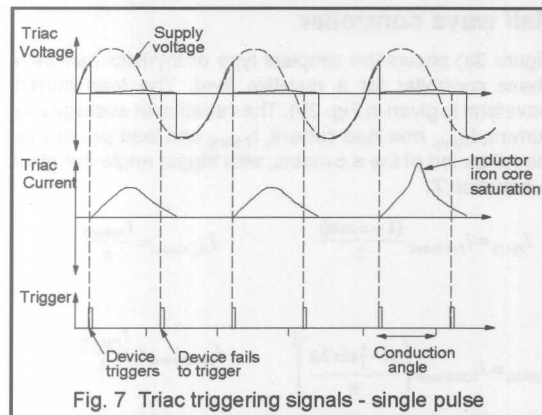


Fig. 7 Triac triggering signals - single pulse

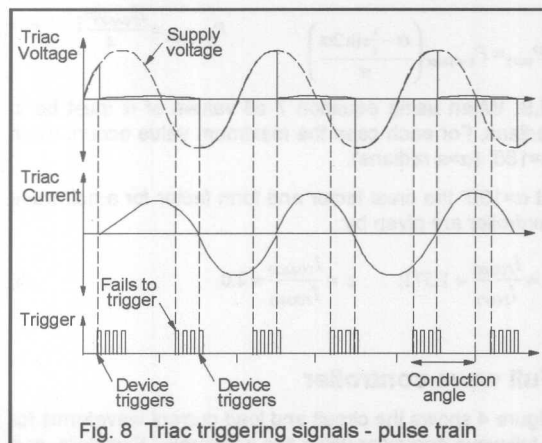
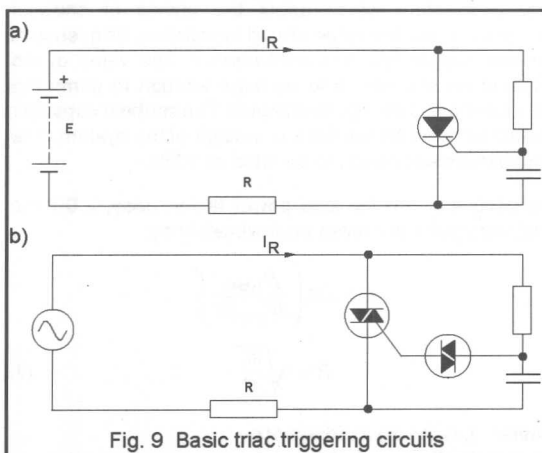
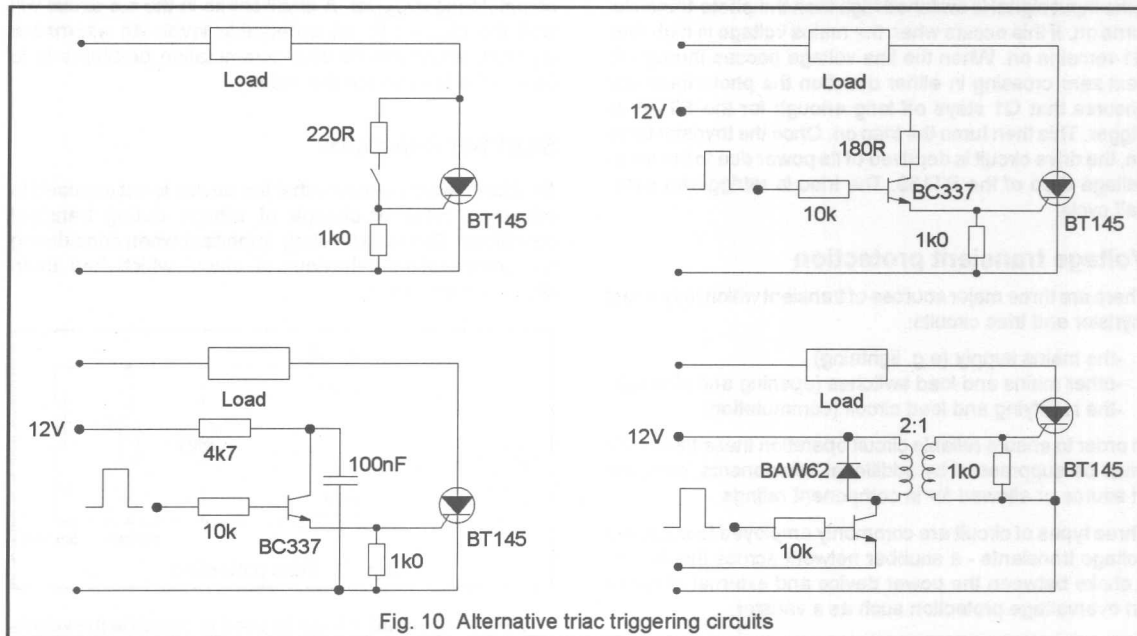


Fig. 8 Triac triggering signals - pulse train

### Gate circuits for thyristors and triacs

As discussed in the introductory article of this chapter, a thyristor or triac can be triggered into conduction when a voltage of the appropriate polarity is applied across the main terminals and a suitable current is applied to the gate. This can be achieved using a delay network of the type shown in Fig. 9a). Greater triggering stability and noise immunity can be achieved if a diac is used (see Fig. 9b). This gives a trigger circuit which is suitable for both thyristors and triacs.

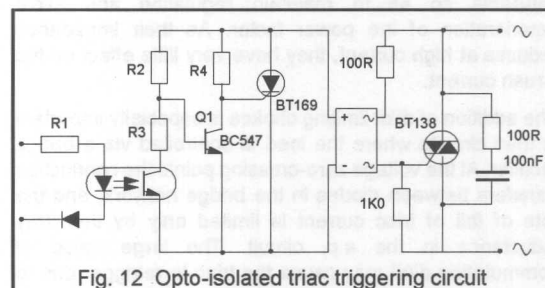
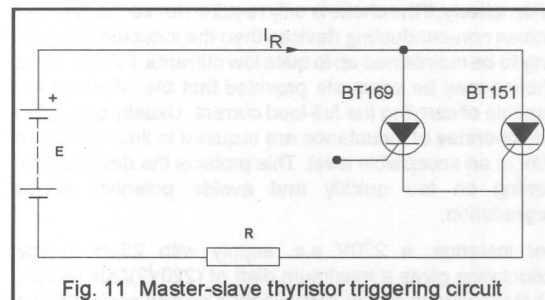
Figure 10 shows several alternative gate drive circuits suitable for typical triac and thyristor applications. In each circuit the gate-cathode resistor protects the device from false triggering due to noise - this is especially important for sensitive gate devices. In addition opto-isolated thyristor and triac drivers are available which are compatible with the Philips range of devices.



In some applications it may be necessary to cascade a sensitive gate device with a larger power device to give a sensitive gate circuit with a high power handling capability. A typical solution which involves triggering the smaller device (BT169) from a logic-level controller to turn on the larger device (BT151) is shown in Fig. 11.

Figure 12 shows an isolated triac triggering circuit suitable for zero voltage switching applications. This type of circuit is also known as a solid state relay (SSR). The function of

the Q1/R2/R3 stage is that the BC547 is on at all instants in time when the applied voltage waveform is high and thus holds the BT169 off. If the BT169 is off then no gate signal is applied to the triac and the load is switched off.



If the input signal is switched high then the photo-transistor turns on. If this occurs when the mains voltage is high then Q1 remains on. When the line voltage passes through its next zero crossing in either direction the photo transistor ensures that Q1 stays off long enough for the BT169 to trigger. This then turns the triac on. Once the thyristor turns on, the drive circuit is deprived of its power due to the lower voltage drop of the BT169. The triac is retriggered every half cycle.

### Voltage transient protection

There are three major sources of transient which may affect thyristor and triac circuits:

- the mains supply (e.g. lightning)
- other mains and load switches (opening and closing)
- the rectifying and load circuit (commutation)

In order to ensure reliable circuit operation these transients must be suppressed by additional components, removed at source or allowed for in component ratings.

Three types of circuit are commonly employed to suppress voltage transients - a snubber network across the device, a choke between the power device and external circuit or an overvoltage protection such as a varistor.

### Series line chokes

A series choke may be used to limit peak fault currents to assist in the fuse protection of thyristors and triacs. If the choke is used in conjunction with fuse protection, it must retain its inductance to very large values of current, and so for this reason it is usually an air-cored component. Alternatively, if the choke is only required to reduce the  $dv/dt$  across non-conducting devices then the inductance needs only to be maintained up to quite low currents. Ferrite-cored chokes may be adequate provided that the windings are capable of carrying the full-load current. Usually only a few microhenries of inductance are required to limit the circuit  $di/dt$  to an acceptable level. This protects the devices from turning on too quickly and avoids potential device degradation.

For instance, a 220V a.c. supply with 20 $\mu$ H source inductance gives a maximum  $di/dt$  of  $(220\sqrt{2})/20 = 16\text{A}/\mu\text{s}$ . Chokes used to soften commutation should preferably be saturable so as to maintain regulation and avoid deterioration of the power factor. As their impedance reduces at high current, they have very little effect on the inrush current.

The addition of  $di/dt$  limiting chokes is especially important in triac circuits where the load is controlled via a bridge rectifier. At the voltage zero-crossing points the conduction transfers between diodes in the bridge network, and the rate of fall of triac current is limited only by the stray inductance in the a.c. circuit. The large value of commutating  $di/dt$  may cause the triac to retrigger due to

commutating  $dv_{(com)}/dt$ . A small choke in the a.c. circuit will limit the  $di_{(com)}/dt$  to an acceptable level. An alternative topology which avoids triac commutation problems is to control the load on the d.c. side.

### Snubber networks

Snubber networks ensure that the device is not exposed to excessive rates of change of voltage during transient conditions. This is particularly important when considering the commutation behaviour of triacs, which has been discussed elsewhere.

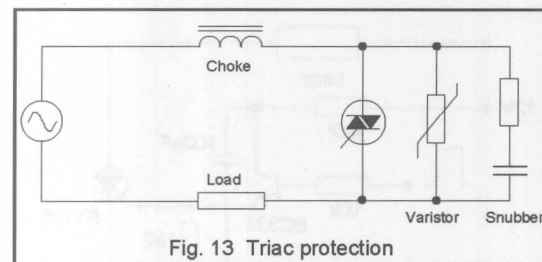


Fig. 13 Triac protection

The following equations can be used to calculate the values of the snubber components required to keep the reapplied  $dv/dt$  for a triac within the  $dv_{(com)}/dt$  rating for that device. The parameters which affect the choice of snubber components are the value of load inductance, frequency of the a.c. supply and rms load current. The value of the snubber resistor needs to be large enough to damp the circuit and avoid voltage overshoots. The snubber capacitor should be rated for the full a.c. voltage of the system. The snubber resistor needs to be rated at 0.5W.

For circuits where the load power factor,  $\cos\phi$ ,  $\geq 0.7$  the snubber values are given approximately by:

$$C \geq 25L \left( \frac{f I_{T(RMS)}}{dv_{(com)}/dt} \right)^2$$

$$R = \sqrt{\frac{3L}{C}} \quad (9)$$

where: L is the load inductance  
f is the supply frequency  
 $I_{T(RMS)}$  is the rms device current  
 $dv_{(com)}/dt$  is the device commutating  $dv/dt$  rating.

The presence of a snubber across the device can improve the turn-on performance of the triac by using the snubber capacitor discharge current in addition to the load current to ensure that the triac latches at turn-on. The value of the snubber resistor must be large enough to limit the peak capacitor discharge current through the triac to within the turn-on  $di/dt$  limit of the device.



### Varistor

The use of a metal oxide varistor (MOV), as shown in Fig. 13, protects the device from transient overvoltages which may occur due to mains disturbances.

### Overcurrent protection

Like all other semiconductor devices, triacs have an infinite life if they are used within their ratings. However, they rapidly overheat when passing excessive current because the thermal capacitance of their junction is small. Overcurrent protective devices (circuit breakers, fuses) must, therefore, be fast-acting.

### Inrush condition

Motors, incandescent lamp or transformer loads give rise to an inrush condition. Lamp and motor inrush currents are avoided by starting the control at a large trigger angle. Transformer inrush currents are avoided by adjusting the initial trigger angle to a value roughly equal to the load phase angle. No damage occurs when the amount of inrush current is below the inrush current rating curve quoted in the device data sheet (see the chapter 'Understanding thyristor and triac data').

### Short-circuit condition

Fuses for protecting triacs should be fast acting, and the amount of fuse  $I^2t$  to clear the circuit must be less than the  $I^2t$  rating of the triac. Because the fuses open the circuit rapidly, they have a current limiting action in the event of a short-circuit. High voltage fuses exhibit low clearing  $I^2t$  but the fuse arc voltage may be dangerous unless triacs with a sufficiently high voltage rating are used.

### Conclusions

This paper has outlined the most common uses and applications of thyristor and triac circuits. The type of circuit used depends upon the degree of control required and the nature of the load. Several types of gate circuit and device protection circuit have been presented. The amount of device protection required will depend upon the conditions imposed on the device by the application circuit. The protection circuits presented here will be suitable for the majority of applications giving a cheap, efficient overall design which uses the device to its full capability with complete protection and confidence.



### 6.1.3 The Peak Current Handling Capability of Thyristors

The ability of a thyristor to withstand peak currents many times the size of its average rating is well known. However, there is little information about the factors affecting the peak current capability. This section will investigate the effect of pulse duration on the peak current capability of thyristors.

Data sheets for thyristors always quote a figure for the maximum surge current that the device can survive. This figure assumes a half sine pulse with a width of either 10 ms or 8.3 ms, which are the conditions applicable for 50/60 Hz mains operation. This limit is not absolute; narrow pulses with much higher peaks can be handled without damage but little information is available to enable the designer to determine how high this current is. This section will discuss some of the factors affecting a thyristor's peak current capability and review the existing prediction methods. It will go on to present the results of an evaluation of the peak current handling capabilities for pulses as narrow as 10  $\mu$ s for the BT151, BT152 and BT145 thyristors. It will also propose a method for estimating a thyristor's peak current capability for a half sine pulse with a duration between 10  $\mu$ s and 10 ms from its quoted surge rating.

#### Energy Handling

In addition to the maximum surge current, data sheets often quote a figure called " $I^2t$  for fusing". This number is used to select appropriate fuses for device protection.  $I^2t$  represents the energy that can be passed by the device without damage. In fact it is not the passage of the energy which causes damage, but the heating of the crystal by the energy absorbed by the device which causes damage.

If the period over which the energy is delivered is long, the absorbed energy has time to spread to all areas of the device capable of storing it - like the edges of the crystal, the plastic encapsulation, the mounting tab and for very long times the heatsink - therefore the temperature rise in the crystal is moderated. If, however, the delivery period is short - say a single half sine pulse of current with a duration of <10 ms - the areas to which the energy can spread for the actual duration of the pulse are limited. This means that the crystal keeps all the energy giving a much bigger temperature rise. For very short pulses (<0.1 ms) and large crystal, the problem is even worse because not all of the active area of a thyristor crystal is turned on simultaneously - conduction tends to spread out from the gate area - so the current pulse passes through only part of the crystal resulting in a higher level of dissipation and an even more restricted area for absorbing it.

#### Expected Results

$I^2t$  is normally quoted at 10 ms, assuming that the surge is a half sine pulse, and is derived from the surge current from:

$$I^2t = \left( \frac{I_{TSM}}{\sqrt{2}} \right)^2 \cdot 0.01$$

This calculates the RMS current by dividing  $I_{TSM}$  by  $\sqrt{2}$

Under the simplest of analyses  $I^2t$  would be assumed to be constant so a device's peak current capability could be calculated from:

$$I_{pk} = I_{TSM} \cdot \left( \frac{0.01}{t_p} \right)^{\frac{1}{2}}$$

where  $I_{pk}$  is the peak of a half sine current pulse with a duration of  $t_p$ . However, experience and experiments have shown that such an approach is inaccurate. To overcome this, other 'rules' have been derived.

One of these 'rules' suggests that it is not  $I^2t$  which is constant but  $I^3t$  or  $I^4t$ . Another suggestion is that the 'constancy' continuously changes from  $I^2t$  to  $I^4t$  as the pulses become shorter. All these rules are expressed in the general equation:

$$I_{pk} = I_{TSM} \cdot \left( \frac{0.01}{t_p} \right)^{\frac{1}{N}}$$

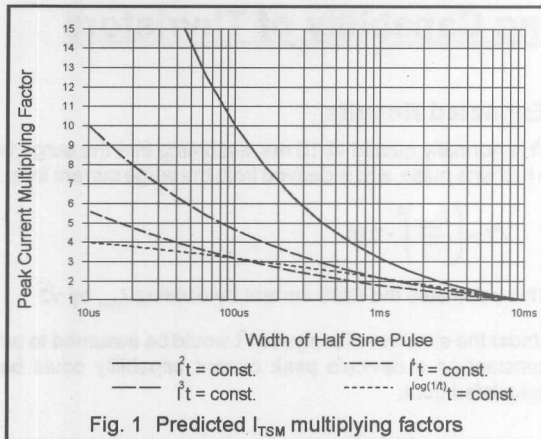
where  $N$  is either constant or a function of the pulse width, for example:

$$N = \log \left( \frac{1}{t_p} \right)$$

The graph shown in Fig. 1 shows what several of these 'rules' predict would happen to the peak current capability if they were true. Unfortunately little or no real information currently exists to indicate the validity of these rules. Tests have been performed on three groups of devices - BT151, BT152 and BT145 - to gather the data which would, hopefully, decide which was correct.

#### Test Circuit

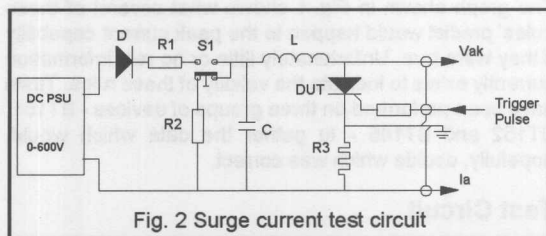
The technique chosen to measure the peak current capability of the devices was the stepped surge method. In this test, the device is subjected to a series of current pulses of increasing magnitude until it receives a surge which causes measurable degradation.



### Circuit Description

The circuits used to perform the required measurements were of the form shown in Fig. 2. They produce half sine pulses of current from the resonant discharge of C via L. Triggering of the device under test (DUT) itself is used to initiate the discharge. The gate signal used for all the tests was a 100 mA / 1 µs pulse fed from a pulse generator in single-shot mode.

The magnitude of the current pulse is adjusted by changing the voltage to which C is initially charged by varying the output of the PSU. The pulse is monitored by viewing the voltage across R3 on an digital storage oscilloscope. R1 and D protect the power supply. R1 limits the current from the supply when DUT fails and during the recharging of C. D attempts to prevent any high voltage spikes being fed back into the PSU.



Pushbutton S1 and resistor R2 are a safety feature. R2 keeps C discharged until S1 is pressed. The trigger pulse needs a button on the pulse generator to be pressed which means both hands are occupied and kept away from the test circuit high voltages.

### Choice of L & C

The width of the half sine pulse from an LC circuit is:

$$t_{pulse} = \pi \cdot \sqrt{L \cdot C}$$

and the theoretical peak value of the current is:

$$I_{peak} = V \cdot \sqrt{\frac{C}{L}}$$

These equations assume that the circuit has no series resistance to damp the resonant action which would result in a longer but lower pulse. Minimising these effects was considered to be important so care was taken during the building of the circuits to keep the resistance to a minimum. To this end capacitors with low ESR were chosen, the inductors were wound using heavy gauge wire and the loop C / L / DUT / R3 was kept as short as possible.

It was decided to test the devices at three different pulse widths - 10 µs, 100 µs and 1 ms - so three sets of L and C were needed. The values were selected with the help of a 'spreadsheet' program running on an PC compatible computer. The values which were finally chosen are shown in Table 1. Also given in Table 1 are the theoretical peak currents that the L / C combination would produce for a initial voltage on C of 600 V.

### Test Procedure

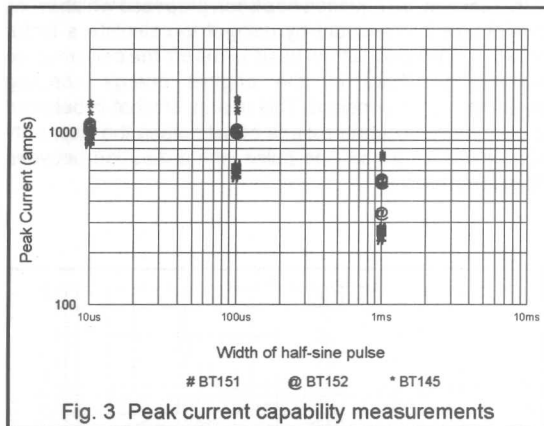
As mentioned earlier, the test method called for each device to be subjected to a series of current pulses of increasing amplitude. The resolution with which the current capability is assessed is defined by the size of each increase in current. It was decided that steps of approximately 5% would give reasonable resolution.

Experimentation indicated that the clearest indication of device damage was obtained by looking for changes in the off-state breakdown voltage. So after each current pulse the DUT was removed from the test circuit and checked on a curve tracer. This procedure did slow the testing but it was felt that it would result in greater accuracy.

Pulse Width	C (µF)	L (µH)	I <sub>peak</sub> (A)
10 µs	13.6	0.75	2564
100 µs	100	10	1885
1 ms	660	154	1244

Table 1. Inductor and Capacitor Values

It was also decided that, since this work was attempting to determine the current that a device could survive - not which killed it, the figure actually quoted in the results for a device's current capability would be the value of the pulse prior to the one which caused damage.



### Test Results

Figure 3 is a graph showing the measured current capabilities of all of the tested devices. Table 2 summarises the measurements by giving the mean of the results for the three device types at each of the pulse widths. Table 3 expresses the mean values as factors of the device  $I_{TSM}$  rating. This table also gives the factors that the various 'rules' would have predicted for the various pulse widths.

Pulse Width	Mean Peak Current Capability (Amps)		
	BT151	BT152	BT145
10 µs	912	1092	1333
100 µs	595	1021	1328
1 ms	264	490	697

Table 2. Measured Current Capability

Pulse Width	Measured Factor			Predicted Factor (by $I^nt$ rule)			
	BT 151	BT 152	BT 145	n=2	n=3	n=4	n=log(1/t)
10 µs	9.1	5.5	4.4	31.6	10.0	5.6	4.0
100 µs	6.0	5.1	4.4	10.0	4.6	3.2	3.2
1 ms	2.6	2.4	2.3	3.2	2.2	1.8	2.2

Table 3. Measured and Predicted  $I_{TSM}$  Multiplication Factors

### Interpretation of Results

It had been hoped that the measurements would give clear indication of which of the 'rules' would give the most accurate prediction of performance. However, an inspection of Table 3 clearly shows that there is no

correlation between any of the predicted factors and the measured factors. In fact the variation in the factors between the various device types would indicate that no rule based on an  $I^nt$  function alone can give an accurate prediction. This implies that something else will have to be taken into account.

Further study of Fig. 3 reveals that the difference in the peak current capability of the three device types is becoming less as the pulses become shorter. This could be explained by a reduction in the active area of the larger crystals, making them appear to be smaller than they actually are. This is consistent with the known fact that not all areas of a thyristor turn on simultaneously - the conduction region tends to spread out from the gate. If the pulse duration is less than the time it takes for all areas of the device to turn on, then the current flows through only part of the crystal, reducing the effective size of the device. If the rate at which the conduction area turns on is constant then the time taken for a small device to be completely ON is shorter than for a large device. This would explain why the performance increase of the BT145 starts falling off before that of the BT151.

### Proposed Prediction Method

The above interpretation leads one to believe that the original energy handling rule, which says that  $I^2t$  is a constant, may still be correct but that the performance it predicts will 'roll off' if the pulse duration is less than some critical value. The equation which was developed to have the necessary characteristics is:

$$I_{pk} = I_{TSM} \cdot \left( \frac{0.01}{t_p} \right)^{\frac{1}{2}} \cdot \left( \frac{t_p}{t_p + t_{crit}} \right)^{\frac{1}{2}}$$

which simplifies to:-

$$I_{pk} = I_{TSM} \cdot \sqrt{\frac{0.01}{t_p + t_{crit}}}$$

where  $t_{crit}$  is proportional to - but not necessarily equal to - the time taken to turn on all the active area of the crystal and is calculated from:-

$$t_{crit} = \frac{A}{R}$$

where: A = crystal area

R = constant expressing the rate at which the area is turned on.

Preferably, A should be the area of the cathode but this information is not always available. As an alternative the total crystal area can be used if the value of R is adjusted accordingly. This will inevitably introduce an error because cathode and crystal areas are not directly proportional, but it should be relatively small.



R was determined empirically to be approximately  $0.02 \text{ m}^2/\text{s}$ . Using this value of R gives the values of  $t_{\text{crit}}$  shown in Table 3. Using these values in the above equation predicts that the peak current handling capability of the BT151, BT152 and BT145 would be as shown in Fig. 4.

Device	$t_{\text{crit}}$
BT151	148 $\mu\text{s}$
BT152	410 $\mu\text{s}$
BT145	563 $\mu\text{s}$

Table 3. Calculated Values of  $t_{\text{crit}}$

### Conclusions

The first conclusion that can be drawn from this work is that a thyristor, with average rating of only 7.5A, is capable of conducting, without damage, a peak current greater than 100 times this value in a short pulse. Furthermore the power required to trigger the device into conducting this current can be  $<1 \mu\text{W}$ . This capability has always been known and indeed the surge rating given in the data sheet gives a value for it at pulse widths of around 10 ms. What has been missing is a reliable method of predicting what the peak current capability of a device is for much shorter pulses.

The results obtained using the test methods indicate that the previously suggested 'rules' fail to take into account the effect that crystal size has on the increase in performance.

In this section, an equation has been proposed which takes crystal size into account by using it to calculate a factor called  $t_{\text{crit}}$ . This time is then used to 'roll off' the performance increase predicted by the original energy handling equation -  $I^2t = \text{constant}$ . This results in what is believed to be a more accurate means of estimating the capability of a device for a half sine pulse with a duration between 10  $\mu\text{s}$  and 10 ms.

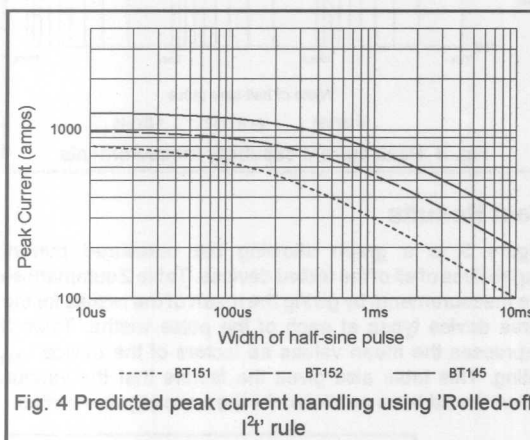


Fig. 4 Predicted peak current handling using 'Rolled-off  $I^2t$ ' rule

### 6.1.4 Understanding Thyristor and Triac Data

The importance of reliable and comprehensive data for power semiconductor devices, together with the advantages of the absolute maximum rating system, is clear. This present article describes the data sheet descriptions of Philips thyristors and triacs, and aims to enable the circuit designer to use our published data to the full and to be confident that it truly describes the performance of the devices.

A brief survey of short-form catalogues is an insufficient method of comparing different devices. Published ratings and characteristics require supporting information to truly describe the capabilities of devices; thus comparisons between devices whose performance appears to be similar should not be made on economic grounds alone. Manufacturers have been known to quote ratings in such a way as to give a false impression of the capabilities of their devices.

Ratings and characteristics given in published data should always be quoted with the conditions to which they apply, and these conditions should be those likely to occur in operation. Furthermore, it is important to define the rating or characteristic being quoted. Only if data is both complete and unambiguous can a true comparison be made between the capabilities of different types.

#### Thyristors

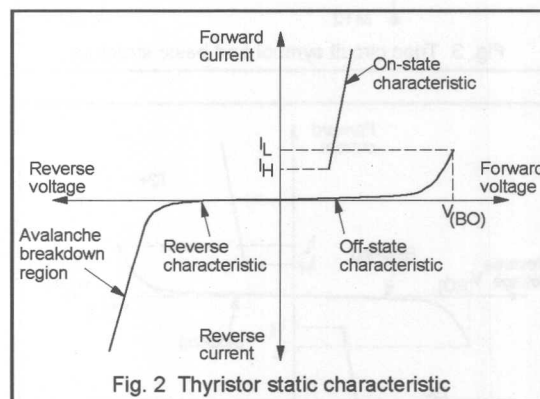
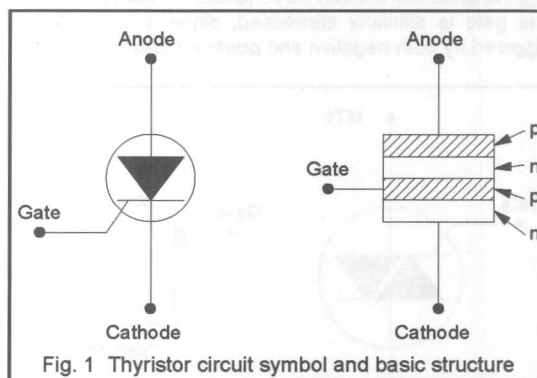
Thyristor is a generic term for a semiconductor device which has four semiconductor layers and operates as a switch, having stable on and off states. A thyristor can have two, three, or four terminals but common usage has confined the term thyristor to three terminal devices. Two-terminal devices are known as switching diodes, and four-terminal devices are known as silicon controlled switches. The common, or three-terminal, thyristor is also known as the reverse blocking triode thyristor or the silicon controlled rectifier (SCR). Fig. 1 shows the circuit symbol and a schematic diagram of the thyristor. All Philips thyristors are p-gate types; that is, the anode is connected to the metal tab.

The thyristor will conduct a load current in one direction only, as will a rectifier diode. However, the thyristor will only conduct this load current when it has been 'triggered'; this is the essential property of the thyristor.

Fig. 2 shows the static characteristic of the thyristor. When a small negative voltage is applied to the device, only a small reverse leakage current flows. As the reverse voltage is increased, the leakage current increases until avalanche breakdown occurs. If a positive voltage is applied, then again a small forward leakage current flows which increases as the forward voltage increases. When the

forward voltage reaches the breakover voltage  $V_{(BO)}$ , turn-on is initiated by avalanche breakdown and the voltage across the thyristor falls to the on state voltage  $V_T$ .

However, turn-on can occur when the forward (anode-to-cathode) voltage is less than  $V_{(BO)}$  if the thyristor is triggered by injecting a pulse of current into the gate. If the device is to remain in the on state, this trigger pulse must remain until the current through the thyristor exceeds the latching current  $I_L$ . Once the on state is established, the holding current  $I_H$  is the minimum current that can flow through the thyristor and still maintain conduction. The load current must be reduced to below  $I_H$  to turn the thyristor off; for instance, by reducing the voltage across the thyristor and load to zero.



Thyristors are normally turned on by triggering with a gate signal but they can also be turned on by exceeding either the forward breakover voltage or the permitted rate of rise



of anode voltage  $dV/dt$ . However, these alternative methods of switching to the conducting state should be avoided by suitable circuit design.

### Triacs

The triac, or bidirectional triode thyristor, is a device that can be used to pass or block current in either direction. It is therefore an a.c. power control device. It is equivalent to two thyristors in anti-parallel with a common gate electrode. However, it only requires one heatsink compared to the two heatsinks required for the anti-parallel thyristor configuration. Thus the triac saves both cost and space in a.c. applications.

Figure 3 shows the triac circuit symbol and a simplified cross-section of the device. The triac has two main terminals MT1 and MT2 (the load connections) and a single gate. The main terminals are connected to both p and n regions since current can be conducted in both directions. The gate is similarly connected, since a triac can be triggered by both negative and positive pulses.

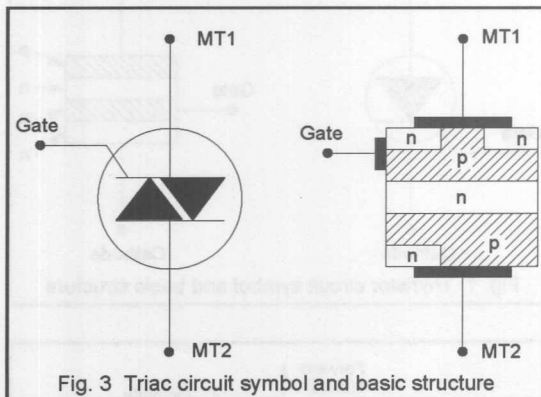


Fig. 3 Triac circuit symbol and basic structure

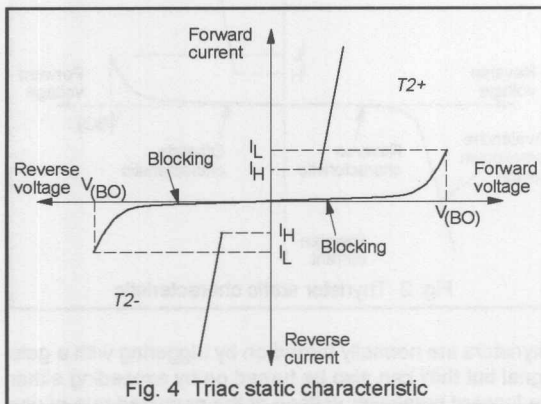


Fig. 4 Triac static characteristic

The on state voltage/current characteristic of a triac resembles that of a thyristor. The triac static characteristic of Fig. 4 shows that the triac is a bidirectional switch. The condition when terminal 2 of the triac is positive with respect to terminal 1 is denoted in data by the term 'T2+'. If the triac is not triggered, the small leakage current increases as the voltage increases until the breakover voltage  $V_{BO}$  is reached and the triac then turns on. As with the thyristor, however, the triac can be triggered below  $V_{BO}$  by a gate pulse, provided that the current through the device exceeds the latching current  $I_L$  before the trigger pulse is removed. The triac, like the thyristor, has holding current values below which conduction cannot be maintained.

When terminal 2 is negative with respect to terminal 1 (T2-) the blocking and conducting characteristics are similar to those in the T2+ condition, but the polarities are reversed. The triac can be triggered in both directions by either negative (G-) or positive (G+) pulses on the gate, as shown in Table 1. The actual values of gate trigger current, holding current and latching current may be slightly different in the different operating quadrants of the triac due to the internal structure of the device.

Quadrant	Polarity of T2 wrt T1	Gate polarity
1 (1+)	T2+	G+
2 (1-)	T2+	G-
3 (3-)	T2-	G-
4 (3+)	T2-	G+

Table 1. Operating quadrants for triacs

### Device data

#### Anode to cathode voltage ratings

The voltage of the a.c. mains is usually regarded as a smooth sinewave. In practice, however, there is a variety of transients, some occurring regularly and others only occasionally (Fig. 5). Although some transients may be removed by filters, thyristors must still handle anode to cathode voltages in excess of the nominal mains value.

The following reverse off-state voltage ratings are given in our published data:

**$V_{RSM}$ :** the non-repetitive peak reverse voltage. This is the allowable peak value of non-repetitive voltage transients, and is quoted with the maximum duration of transient that can be handled (usually  $t < 10\text{ms}$ ).

**$V_{RRM}$ :** the repetitive peak reverse voltage. This is the allowable peak value of transients occurring every cycle.

**$V_{RWM}$ :** the peak working reverse voltage. This is the maximum continuous peak voltage rating in the reverse direction, neglecting transients. It corresponds to the peak negative value (often with a safety factor) of the sinusoidal supply voltage.

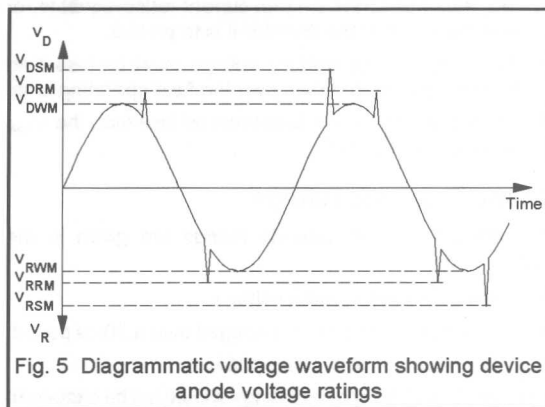


Fig. 5 Diagrammatic voltage waveform showing device anode voltage ratings

The forward off-state voltages corresponding to  $V_{RSM}$ ,  $V_{RRM}$  and  $V_{RWM}$  are listed below.

**$V_{DSM}$ :** the non-repetitive peak off-state voltage applied in the forward direction.

**$V_{DRM}$ :** the repetitive peak off-state voltage applied in the forward direction.

**$V_{DWM}$ :** the peak working off-state voltage applied in the forward direction.

Both the repetitive and non-repetitive voltage ratings are determined partly by the voltage limit that prevents the thyristor being driven into forward or reverse breakdown, and partly by the instantaneous energy (resulting from an increase in leakage current) that can be dissipated in the device without exceeding the rated junction temperature.

When a thyristor is to operate directly from the mains supply, it is advisable to choose a device whose repetitive peak voltage ratings  $V_{RRM}$  and  $V_{DRM}$  are at least 1.5 times the peak value of the sinusoidal supply voltage. This figure forms part of the device type number; for example BT151-650R, where 650 corresponds to  $V_{DRM}$ ,  $V_{RRM}=650V$  and the final R (for Reverse) indicates that the anode of the device is connected to the metal tab.

### Anode-to-cathode current ratings

The following current ratings, described by the waveforms shown in Fig. 6, are given in our published data. Note that the suffix  $_T$  implies that the thyristor is in the on state.

**$I_{T(AV)}$ :** the average value of the idealised mains current waveform taken over one cycle, assuming conduction over  $180^\circ$ . For devices mounted on heatsinks, the  $I_{T(AV)}$  rating should be quoted for a particular mounting-base temperature  $T_{mb}$ ; our devices are generally characterised at a mounting-base temperature of at least  $85^\circ C$ . A device can have an artificially high current rating if the

mounting-base temperature is unrealistically low; ratings with no associated mounting-base temperature should be regarded with suspicion.

**$I_{T(RMS)}$ :** the rms on-state current. This rating gives the maximum rms current that the thyristor can handle. It is important for applications when the device current waveform is described by a high value form factor. For such conditions the rms current rather than the average current may be the limiting rating.

**$I_{TRM}$ :** the repetitive peak forward current. This rating is the peak current that can be drawn each cycle providing that the average and rms current ratings are not exceeded.

**$I_{TSM}$ :** the non-repetitive (surge) peak forward current. This rating is the peak permitted value of non-repetitive transients, and depends on the duration of the surge. Our published data quotes the  $I_{TSM}$  rating for  $t=10ms$ , the duration of a half-cycle of 50Hz mains. However, some manufacturers quote  $I_{TSM}$  for  $t=8.3ms$  (half-cycle of 60Hz mains), and thus surge ratings for devices quoted at  $t=8.3ms$  should be approximately downrated (multiplied by 0.83) before comparing them with  $t=10ms$  surge ratings.

The surge rating also depends on the conditions under which it occurs. Our data sheets quote  $I_{TSM}$  rating under the worst probable conditions, that is,  $T_J=T_{J(max)}$  immediately prior to the surge, followed by reapplied  $V_{RWM(max)}$  immediately after the surge. An unrealistically high  $I_{TSM}$  rating could be quoted if, for example,  $T_J < T_{J(max)}$  prior to the surge and then the full rated voltage is not reapplied.

Published data also includes curves for  $I_{TSM}$  against time which show the maximum allowable rms current which can occur during inrush or start-up conditions. The duration of the inrush transient and the mounting base temperature prior to operation determine the maximum allowable rms inrush current.

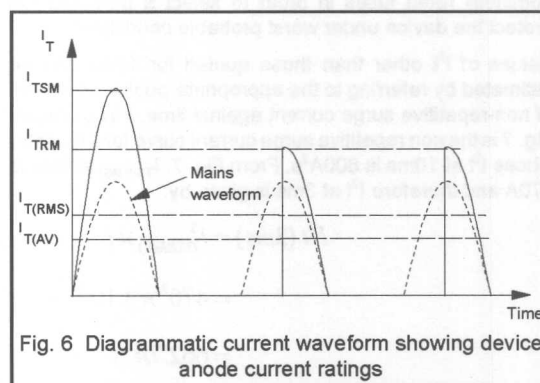


Fig. 6 Diagrammatic current waveform showing device anode current ratings

**di/dt:** the rate of rise of on-state current permissible after triggering. An excessive rate of rise of current causes local heating and thus damage to the device. The rate of rise of current is determined by both the supply and load impedances, and can be limited by additional series inductance in the circuit.

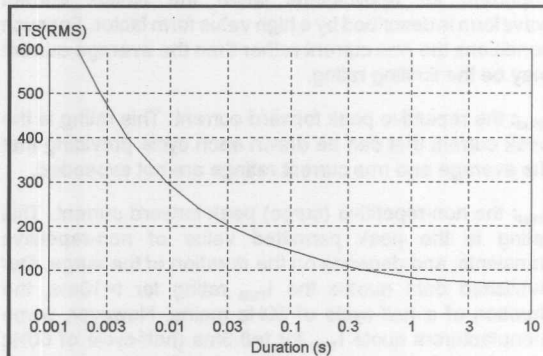


Fig. 7 Non-repetitive surge current as a function of time

**$I^2t$ :** a dimensional convenience specifying the capability of a thyristor to absorb energy. This rating is required for the selection of fuses to protect the thyristor against excessive currents caused by fault conditions. It is normally only valid over the range 3 to 10ms. In our published data, a value is quoted for 10ms, in which case:

$$I^2t = \int i^2 \cdot dt \quad (1)$$

$$= \left( \frac{I_{TSM}}{\sqrt{2}} \right)^2 \times 10 \cdot 10^{-3} \quad (A^2s)$$

The user should match the minimum  $I^2t$  capability of the thyristor to the worst case  $I^2t$  let-through of a range of nominally rated fuses in order to select a fuse that will protect the device under worst probable conditions.

Values of  $I^2t$  other than those quoted for 10ms can be estimated by referring to the appropriate published curves of non-repetitive surge current against time. For example, Fig. 7 is the non repetitive surge current curve for a thyristor whose  $I^2t$  at 10ms is  $800A^2s$ . From Fig. 7,  $I_{TS(RMS)}$  at 3ms is 470A and therefore  $I^2t$  at 3ms is given by:

$$I^2t (3ms) = I_{TS(RMS)}^2 \times t$$

$$= 470^2 \times 3 \cdot 10^{-3}$$

$$= 662.7 A^2s$$

To summarise, when selecting an appropriate fuse the following conditions must be taken into account.

1. The fuse must have an rms current rating equal to, or less than, that of the thyristor it is to protect.
2. The  $I^2t$  at the rms working voltage must be less than that of the thyristor taken over the fuse operating time.
3. The arc voltage of the fuse must be less than the  $V_{RSM}$  rating of the thyristor.

### Gate-to-cathode ratings

The following gate-to-cathode ratings are given in the published data.

**$V_{RGM}$ :** the gate peak reverse voltage.

**$P_{G(AV)}$ :** the mean gate power, averaged over a 20ms period.

**$P_{GM}$ :** the peak gate power dissipation.

The gate-to-cathode power ratings should not be exceeded if over-heating of the gate-cathode junction is to be avoided.

### Temperature ratings

Two temperature ratings are given in the published data.

**$T_{stg}$ :** the storage temperature. Both maximum and minimum values of the temperature at which a device can be stored are given.

**$T_J$ :** the junction temperature. This is one of the principal semiconductor ratings since it limits the maximum power that a device can handle. The junction temperature rating quoted in our published data is the highest value of junction temperature at which the device may be continuously operated to ensure a long life.

### Thermal characteristics

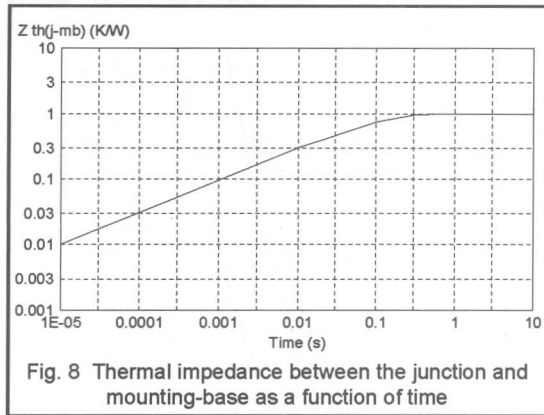
The following thermal resistances and impedances are given in our data.

**$R_{th(j-a)}$ :** the thermal resistance between the junction of the device and ambient (assumed to be the surrounding air).

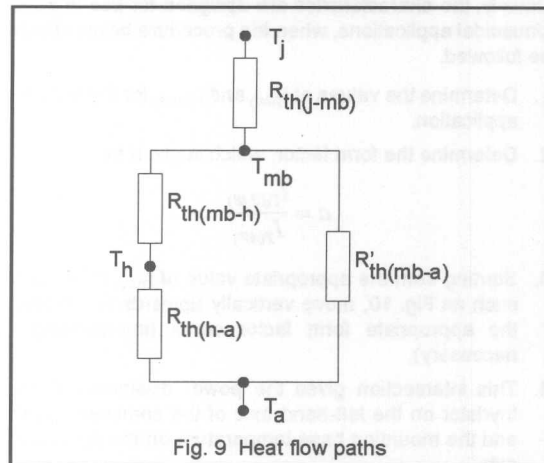
**$R_{th(j-mb)}$ :** the thermal resistance between the junction and mounting base of the device.

**$R_{th(mb-h)}$ :** the thermal resistance between the mounting base of the device and the heatsink (contact thermal resistance).

**$Z_{th(j-mb)}$ :** the transient thermal impedance between the junction and mounting-base of the device. The value given in the published data is for non-repetitive conditions and a particular pulse duration. Under pulse conditions, thermal impedances rather than thermal resistances should be considered. Higher peak power dissipation is permitted under pulse conditions since the materials in a thyristor have a definite thermal capacity, and thus the critical junction temperature will not be reached instantaneously, even when excessive power is being dissipated in the device. The published data also contains graphs of  $Z_{th(j-mb)}$  against time (for non-repetitive conditions) such as those shown in Fig. 8.

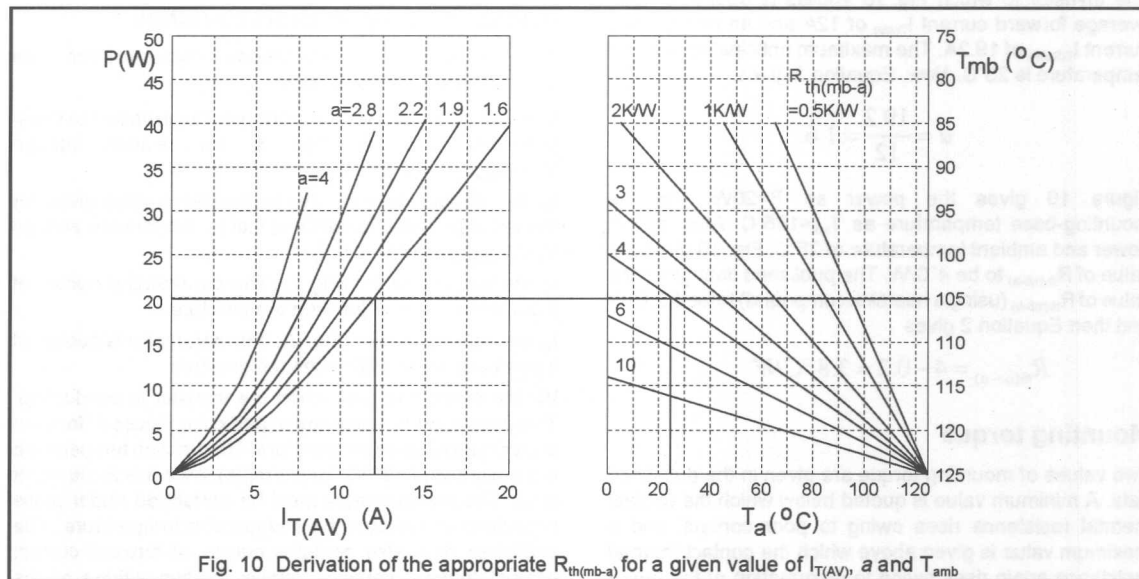


The values of the various thermal resistances between the thyristor junction and the surroundings must be considered to ensure that the junction temperature rating is not exceeded. The heat generated in a semiconductor chip flows by various paths to the surroundings. Fig. 9 shows the various thermal resistances to be taken into account in this process. With no heatsink, the thermal resistance from the mounting-base to the surroundings is given by  $R_{th(mb-a)}$ . When a heatsink is used, the heat loss direct to the surroundings from the mounting-base is negligible owing to the relatively high value of  $R_{th(mb-a)}$  and thus:



$$R_{th(mb-a)} = R_{th(mb-h)} + R_{th(h-a)} \quad (2)$$

Where appropriate, our published data contains power graphs such as that in Fig. 10. These characteristics relate the total power  $P$  dissipated in the thyristor, the average forward current  $I_{T(AV)}$ , the ambient temperature  $T_a$ , and the thermal resistance  $R_{th(mb-a)}$ , with the form factor,  $a$ , as a parameter. They enable the designer to work out the required mounting arrangement from the conditions under which the thyristor is to be operated.





Usually, the characteristics are designed for use in 50Hz sinusoidal applications, when the procedure below should be followed.

1. Determine the values of  $I_{T(AV)}$  and  $I_{T(RMS)}$  for the relevant application.
2. Determine the form factor, which is given by:

$$a = \frac{I_{T(RMS)}}{I_{T(AV)}} \quad (3)$$

3. Starting from the appropriate value of  $I_{T(AV)}$  on a curve such as Fig. 10, move vertically upwards to intersect the appropriate form factor curve (interpolating if necessary).
4. This intersection gives the power dissipated in the thyristor on the left-hand axis of the combined graph and the mounting base temperature on the right hand axis.
5. Moving horizontally across from this intersection to the appropriate value of ambient temperature gives the required mounting base to ambient thermal resistance  $R_{th(mb-a)}$ .
6. The required heatsink thermal resistance  $R_{th(h-a)}$  can now be calculated from Equation 2 since the mounting base to heatsink thermal resistance  $R_{th(mb-h)}$  is given in the published data.

#### Example

The thyristor to which Fig. 10 applies is operated at an average forward current  $I_{T(AV)}$  of 12A and an rms forward current  $I_{T(RMS)}$  of 19.2A. The maximum anticipated ambient temperature is 25°C. Now, Equation 3 gives,

$$a = \frac{19.2}{12} = 1.6$$

Figure 10 gives the power as  $P=20W$  and the mounting-base temperature as  $T_{mb}=105^\circ C$ . Also, at this power and ambient temperature of 25°C, Fig. 10 gives the value of  $R_{th(mb-a)}$  to be 4°C/W. The published data gives the value of  $R_{th(mb-h)}$  (using a heatsink compound) to be 0.2°C/W and then Equation 2 gives

$$R_{th(h-a)} = 4 - 0.2 = 3.8^\circ C/W$$

#### Mounting torque

Two values of mounting torque are given in the published data. A minimum value is quoted below which the contact thermal resistance rises owing to poor contact, and a maximum value is given above which the contact thermal resistance again rises owing to deformation of the tab or cracking of the crystal.

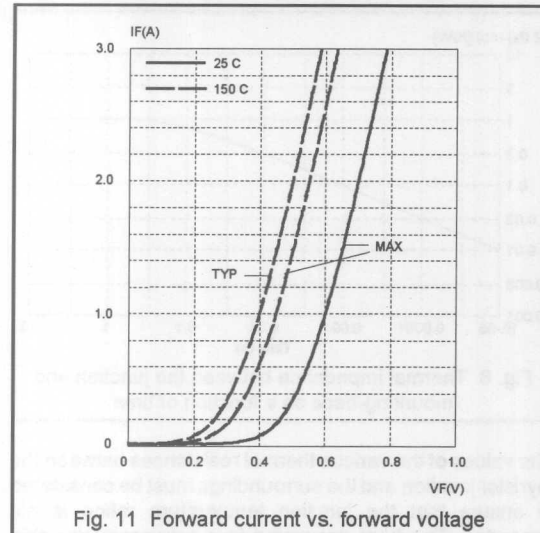


Fig. 11 Forward current vs. forward voltage

The surface of a device case and heatsink cannot be perfectly flat, and thus contact will take place on several points only, with a small air-gap over the rest of the contact area. The use of a soft substance to fill this gap will lower the contact thermal resistance. We recommend the use of proprietary heatsinking compounds which consist of a silicone grease loaded with an electrically insulating and good thermal conducting powder such as alumina.

#### Anode-to-cathode characteristics

The following anode-to-cathode characteristics are included in the published data.

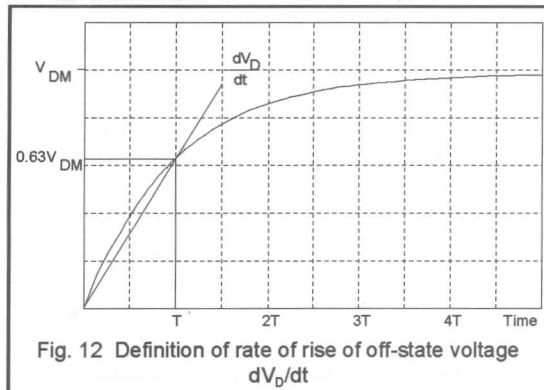
$I_R$ : the reverse current. This parameter is given for the worst probable conditions; that is, the reverse voltage  $V_R = V_{RWM(max)}$  and a high  $T_j$ .

$I_O$ : the off-state current. This parameter is again given for the worst probable conditions; that is, the forward voltage  $V_D = V_{DWM(max)}$  and a high  $T_j$ .

$I_L$ : the latching current (Fig. 2). This parameter is quoted at a particular value of junction temperature.

$I_H$ : the holding current (Fig. 2). This parameter is quoted at a particular value of junction temperature.

$V_T$ : the forward voltage when the thyristor is conducting. This parameter is measured at particular values of forward current and junction temperature. The junction temperature is usually low ( $T_j=25^\circ C$ , for example) since this is the worst case. The measurement must be performed under pulse conditions to maintain the low junction temperature. The published data also contains curves of forward current against forward voltage, usually for two values of the junction temperature: 25°C and  $T_{j(max)}$  (Fig. 11).



$dV/dt$ : the rate of rise of off-state voltage that will not trigger any device. This parameter is given at maximum values of junction temperature  $T_{j(max)}$  and forward voltage  $V_D = V_{DRM(max)}$ .

The values of  $dV_D/dt$  quoted in our published data are normally specified assuming an exponential waveform. This facilitates the design of RC snubber circuits for device protection when required. Fig. 12 illustrates the definition of  $dV_D/dt$ . The final voltage applied to the device  $V_{DM}$  is chosen as  $V_{DRM(max)}$  and the junction temperature is  $T_j = T_{j(max)}$ .

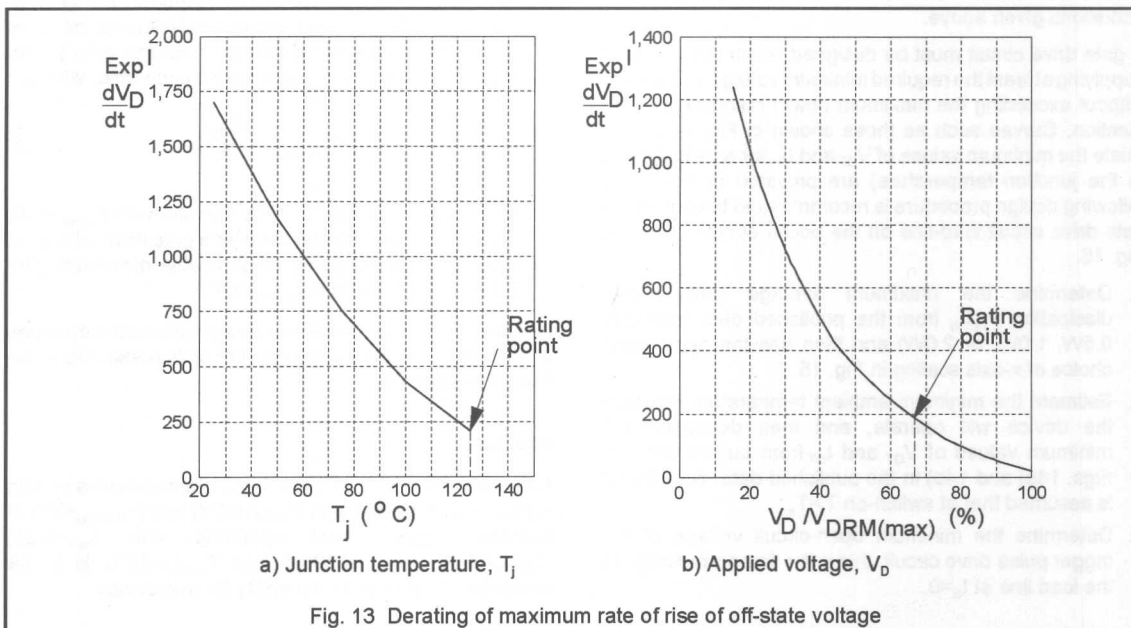
Fig. 12 shows that  $dV_D/dt$  is given by the expression:

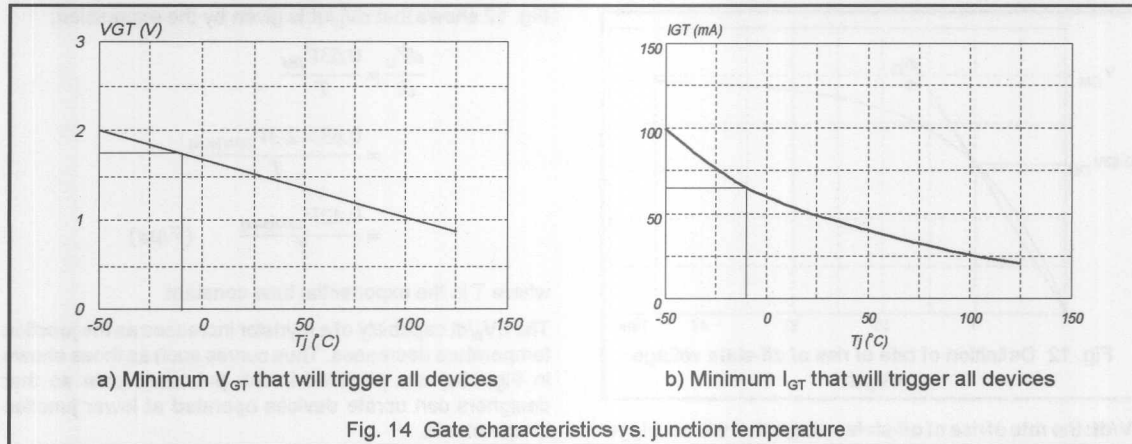
$$\begin{aligned} \frac{dV_D}{dt} &= \frac{0.63V_{DM}}{T} \\ &= \frac{0.63 \times 2/3 V_{DRM(max)}}{T} \\ &= \frac{0.42V_{DRM(max)}}{T} \quad (V/\mu s) \end{aligned}$$

where  $T$  is the exponential time constant.

The  $dV_D/dt$  capability of a thyristor increases as the junction temperature decreases. Thus curves such as those shown in Fig. 13a) are provided in the published data so that designers can uprate devices operated at lower junction temperatures.

The  $dV_D/dt$  characteristic can also be increased by operating the device at a low supply voltage. Thus the published data also contains curves such as Fig. 13b) which shows how  $dV_D/dt$  increases as the ratio  $V_{DM}/V_{DRM(max)}$  decreases. Note that  $V_{DM}$  is unlikely to be greater than  $2/3 V_{DRM(max)}$  (usually owing to the restriction of  $V_{DWM(max)}$ ) and therefore the fact that  $dV_D/dt$  approaches zero as  $V_{DM}$  increases above the value of  $2/3 V_{DRM(max)}$  does not cause problems.





### Gate-to-cathode characteristics

The following gate-to-cathode characteristics are given in the published data.

$V_{GT}$ : the gate-to-cathode voltage that will trigger all devices. This characteristic should be quoted for particular values of applied voltage  $V_D$  and low junction temperature.

$I_{GT}$ : the gate-to-cathode current that will trigger all devices. This characteristic should be quoted for the same conditions given above.

A gate drive circuit must be designed which is capable of supplying at least the required minimum voltage and current without exceeding the maximum power rating of the gate junction. Curves such as those shown in Fig. 14 (which relate the minimum values of  $V_{GT}$  and  $I_{GT}$  for safe triggering to the junction temperature) are provided in data. The following design procedure is recommended to construct a gate drive circuit load-line on the power curves shown in Fig. 15.

1. Determine the maximum average gate power dissipation  $P_{G(AV)}$  from the published data (normally 0.5W, 1.0W, or 2.0W) and then use the appropriate choice of x-axis scaling in Fig. 15.
2. Estimate the minimum ambient temperature at which the device will operate, and then determine the minimum values of  $V_{GT}$  and  $I_{GT}$  from curves such as Figs. 14a) and 14b) in the published data. Note that it is assumed that at switch-on  $T_J = T_a$ .
3. Determine the minimum open-circuit voltage of the trigger pulse drive circuit: this is the first co-ordinate on the load line at  $I_G = 0$ .

4. Using the appropriate horizontal scaling for the device ( $P_{G(AV)} = 0.5W, 1.0W$  or  $2.0W$ ), plot a second point on the power curve whose co-ordinates are given by  $V_{GT(min)}$  and  $5 \times I_{GT(min)}$ . Construct a load line between these two points. The slope of this load gives the maximum allowable source resistance for the drive circuit.
5. Check the power dissipation by ensuring that the load line must not intersect the curve for the maximum peak gate power  $P_{GM(max)}$  which is the outermost ( $\delta = 0.1$ ) curve of Fig. 15. The load line must also not intersect the curve which represents the maximum average gate power  $P_{G(AV)}$  modified by the pulse mark-space ratio, where:

$$P_{GM(max)} = \frac{P_{AV}}{\delta} \quad (5)$$

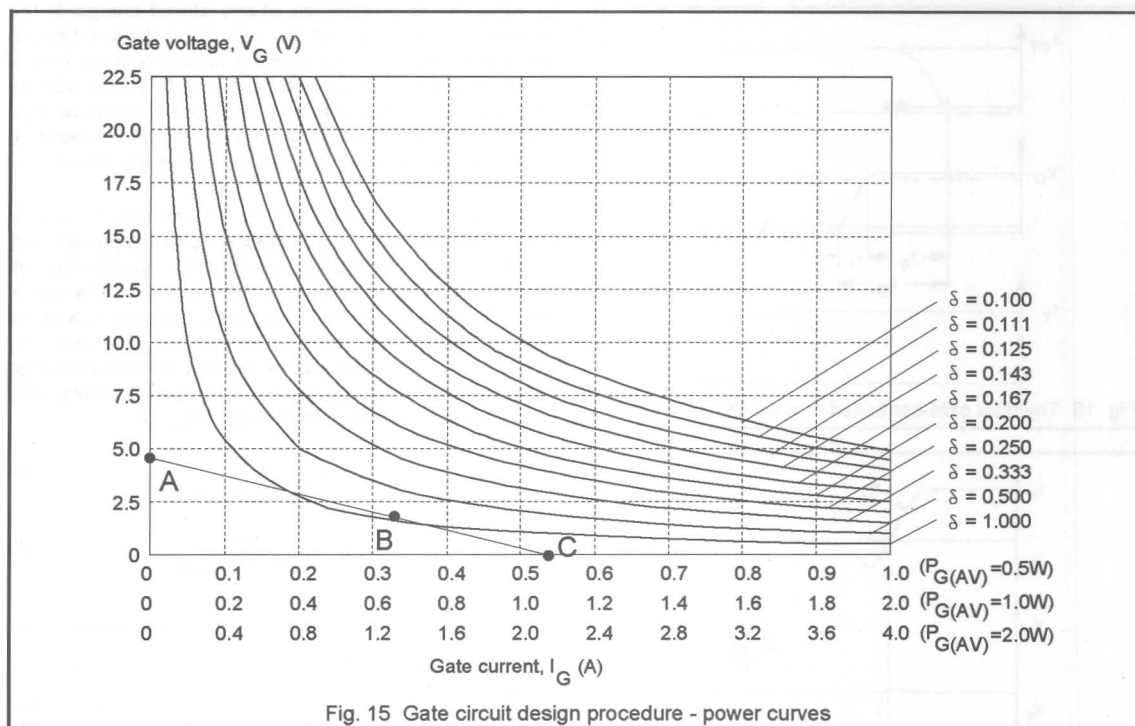
For instance, in Fig. 15, for a thyristor with  $P_{G(AV)} = 1W$ , the  $\delta = 0.25$  curve can be used for a gate drive with a 1:3 mark-space ratio giving an allowable maximum gate power dissipation of  $P_{GM(max)} = 4W$ .

An illustration of how the above design procedure operates to give an acceptable gate drive circuit is presented in the following example.

### Example

A thyristor has the  $V_{GT}/T_J$  and  $I_{GT}/T_J$  characteristics shown in Fig. 14 and is rated with  $P_{G(AV)} = 0.5W$  and  $P_{GM(max)} = 5W$ . A suitable trigger circuit operating with  $\delta_{max} = 0.25$ ,  $V_{GT(min)} = 4.5V$ ,  $I_{GT(max)} = 620mA$  and  $T_{a(min)} = -10^\circ C$  is to be designed. Determine its suitability for this device.





1. Select the top x-axis scale of Fig. 15 ( $P_{G(AV)} = 0.5W$ ).
2. From Fig. 14,  $V_{GT(min)} = 1.75V$ , and  $I_{GT(min)} = 66mA$ .
3. At minimum supply voltage, the open-circuit gate voltage is 4.5V, giving point 'A' in Fig. 15. Point B is plotted at the co-ordinates  $V_{GT(min)}$  and  $5I_{GT(min)}$ , that is at 1.75V and 330mA, and load line ABC is constructed as shown. Note that point C is the maximum current required at  $I_G = 570mA$  and is within the capability of the drive circuit.
4. As required the load line does not intersect the  $P_{G(max)}$  ( $\delta = 0.1$ ). The gate drive duty cycle,  $\delta$ , is 0.25. Therefore  $P_{GM(max)} = P_{G(AV)} / \delta = 0.5 / 0.25 = 2W$ . As required, the load line ABC does not intersect the  $\delta = 0.25$  curve.

### Switching characteristics

Two important switching characteristics are usually included in our published data. They are the gate-controlled turn-on time  $t_{gt}$  (divided into a turn-on delay time,  $t_d$ , and a rise time,  $t_r$ ) and the circuit-commutated turn-off time,  $t_q$ .

#### Gate-controlled turn-on time, $t_{gt}$

Anode current does not commence flowing in the thyristor at the instant that the gate current is applied. There is a period which elapses between the application of the trigger

pulse and the onset of the anode current which is known as the delay time  $t_d$  (Fig. 16). The time taken for the anode voltage to fall from 90% to 10% of its initial value is known as the rise time  $t_r$ . The sum of the delay time and the rise time is known as the gate-controlled turn-on time  $t_{gt}$ .

The gate controlled turn-on time depends on the conditions under which it is measured, and thus the following conditions should be specified in the published data.

- Off-state voltage; usually  $V_D = V_{DWM(max)}$ .
- On-state current.
- Gate trigger current; high gate currents reduce  $t_{gt}$ .
- Rate of rise of gate current; high values reduce  $t_{gt}$ .
- Junction temperature; high temperatures reduce  $t_{gt}$ .

#### Circuit-commutated turn-off time

When a thyristor has been conducting and is reverse-biased, it does not immediately go into the forward blocking state: minority charge carriers have to be cleared away by recombination and diffusion processes before the device can block reapplied off-state voltage. The time from the instant that the anode current passes through zero to the instant that the thyristor is capable of blocking reapplied off-state voltage is the circuit-commutated turn-off time  $t_q$  (Fig. 17).

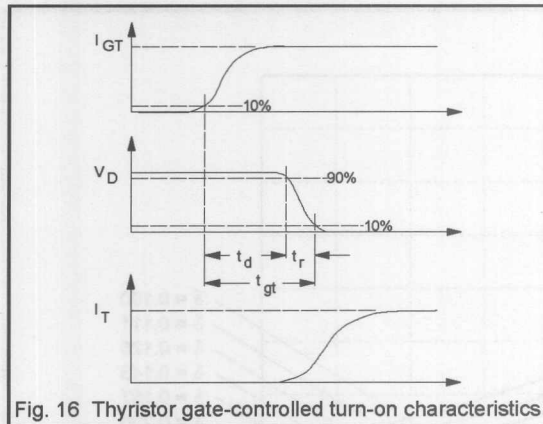


Fig. 16 Thyristor gate-controlled turn-on characteristics

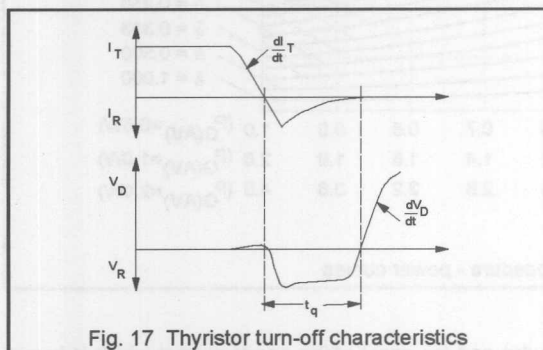


Fig. 17 Thyristor turn-off characteristics

The following conditions should be given when  $t_q$  is quoted.

- On-state current; high currents increase  $t_q$ .
- Reverse voltage; low voltages increase  $t_q$ .
- Rate of fall of anode current; high rates increase  $t_q$ .
- Rate of rise of re-applied off-state voltage; high rates increase  $t_q$ .
- Junction temperature; high temperatures increase  $t_q$ .
- Gate bias; negative voltages decrease  $t_q$ .

### Triac ratings

The ratings and characteristics of the triac are similar to those of the thyristor, except that the triac does not have any reverse voltage ratings (a reverse voltage in one quadrant is the forward voltage in the opposite quadrant). However, one characteristic requires special attention when choosing triacs; the rate of re-applied voltage that the triac will withstand without uncontrolled turn-on.

If a triac is turned off by simply rapidly reversing the supply voltage, the recovery current in the device would simply switch it on in the opposite direction. To guarantee reduction of the current below its holding value, the supply voltage must be reduced to zero and held there for a sufficient time

to allow the recombination of any stored charge in the device. To ensure turn-off, the rate of fall of current during the commutation interval (turn-off period) and the rate of rise of re-applied voltage after commutation must both be restricted. An excessive rate of fall of current creates a large number of residual charge carriers which are then available to initiate turn-on when the voltage across the triac rises.

With supply frequencies up to around 400Hz and a sinusoidal waveform, commutation does not present any problems when the load is purely resistive, since the current and voltage are in phase. As shown in Fig. 18 the rate of fall of on-state current  $-di/dt$ , given by Equation 6, and the rate of rise of commutating voltage  $dV_{com}/dt$ , given by equation 7, are sufficiently low to allow the stored charge in the device to fully recombine. The triac is thus easily able to block the rising re-applied voltage  $dV_{com}/dt$ .

$$di/dt = 2\pi f \cdot \sqrt{2} I_{T(RMS)} \quad (6)$$

$$dV_{com}/dt = 2\pi f \cdot \sqrt{2} V_{(RMS)} \quad (7)$$

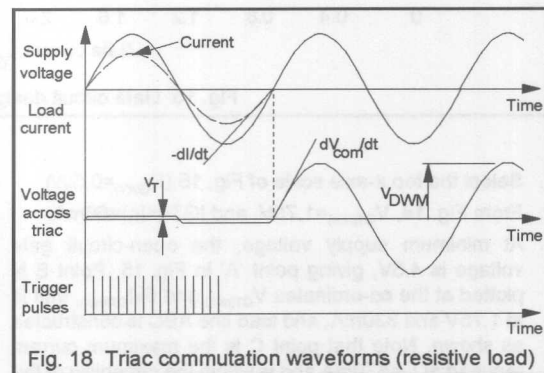


Fig. 18 Triac commutation waveforms (resistive load)

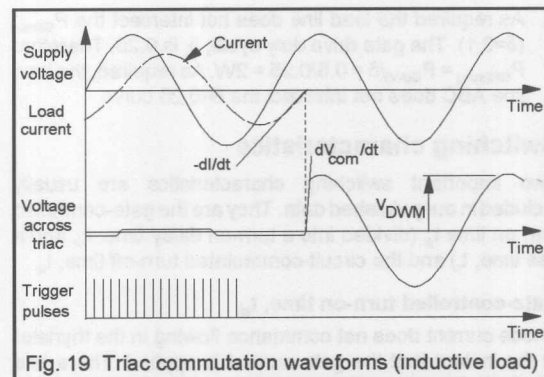
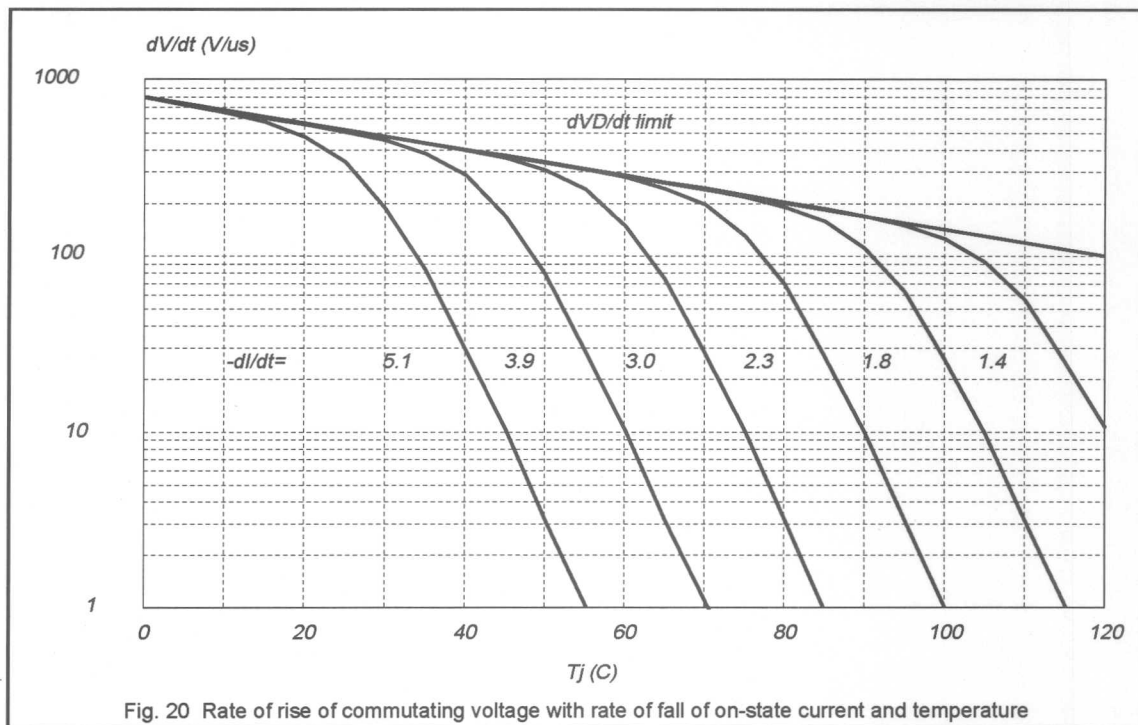


Fig. 19 Triac commutation waveforms (inductive load)



However, with an inductive load (Fig. 19) the current lags behind the voltage and consequently commutation can present special difficulties. When the on-state current has fallen to zero after a triac has been conducting in one direction the supply voltage in the opposite direction will have already reached a significant value. The rate of fall of triac current will still be given by Equation 6 but the rate of rise of reapplied voltage,  $dV_{com}/dt$  will be very large. The triac may switch on immediately unless  $dV/dt$  is held less than that quoted in the published data by suitable circuit design. Alternatively, the circuit design can remain simple if Hi-Com triacs are employed instead. Sections 6.3.1 and 6.3.2 explain the advantages of using Hi-Com triacs in such inductive circuits.

The maximum rate of rise of commutating voltage which will not cause the device to trigger spuriously is an essential part of the triac published data. However,  $dV_{com}/dt$  is meaningless unless the conditions which are applicable are provided, particularly the rate of fall of on-state current  $-di_T/dt$ . Our published data also contains graphs such as

Fig. 20 which relate  $dV_{com}/dt$  to junction temperature with  $-di_T/dt$  as a parameter. The characteristic  $dV_{com}/dt$  is specified under the worst probable conditions, namely:

- mounting base temperature,  $T_{mb} = T_{mb(max)}$
- reapplied off-state voltage,  $V_D = V_{DWM(max)}$
- rms current,  $I_{T(RMS)} = I_{T(RMS)(max)}$

In order that designers may economise their circuits as far as possible, we offer device selections with the same current ratings but with different values of  $dV_{com}/dt$  (at the same value of  $-di_T/dt$ ) for some of our triac families. The  $dV/dt$  capability can be traded off against the gate sensitivity ( $I_{GT(max)}$ ) of the device. Sensitive gate triacs (i.e. those which require only a small amount of gate current to trigger the device) have less ability to withstand high values of  $dV_{com}/dt$  before sufficient current flows within the device to initiate turn-on. These different device selections are differentiated by suffixes which are added to the device type number eg. BT137-600F.

Detailed design considerations for  $dV_{com}/dt$  limiting in inductive circuits when using triacs are considered in separate articles in this handbook.





*Thyristor and Triac Applications*



## 6.2.1 Triac Control of DC Inductive Loads

### The problem of inductive loads

This publication investigates the commutation problem encountered when triacs are used in phase control circuits with inductive loads. Commutation failure is likely to occur owing to circuit inductance imposing a sudden rise of voltage on the triac after conduction. Control of transformers supplying an inductively loaded bridge rectifier is particularly troublesome because of the added effect of rapid current decay during commutation. For a better understanding of the nature of the problem, the commutation behaviour is summarised here.

Triacs are bipolar power control elements that may turn on with either polarity of voltage applied between their main terminals. Unlike thyristors there is no circuit-imposed turn-off time. To ensure commutation the decay rate of current before turn-off and the rate of rise of reapplied voltage must both be held below specified limits. An excessive current decay rate has a profound effect on the maximum rate of rise of voltage that can be sustained, as then a large amount of stored charge is available to initiate the turn-on in the next half cycle.

Figure 1 shows the condition for a triac controlled transformer followed by a rectifier with inductive load. The load inductance forces the rectifier diodes into conduction whenever the instantaneous dc output voltage drops to zero. The transformer secondary is thus shorted for some time after the zero transitions of the mains voltage and a reverse voltage is applied to the triac, turning it off. Because of transformer leakage inductance the triac does not turn off immediately but continues to conduct over what is called the commutation interval (see Fig. 1).

During the commutation interval a high rate of decay of current ( $di_{com}/dt$ ) results for two reasons. Firstly the rate of fall of current is high because the leakage inductance of most transformers is low. This is necessary to achieve a small dc output voltage loss (represented by the shaded areas in the voltage waveform of Fig. 1) in the transformer. Secondly, with an inductive rectifier load a substantial current flows when commutation starts to occur.

The large value of  $di_{com}/dt$  results in a high rate of rise of voltage,  $dv/dt$ . Since the current decays rapidly the peak reverse recovery current  $I_{RRM}$  is fairly large. Upon turn-off,  $I_{RRM}$  is abruptly transferred to the snubber elements R and C so the voltage abruptly rises to the level  $R \cdot I_{RRM}$  (C is initially discharged). Owing to the high value of both  $di_{com}/dt$  and  $dv/dt$ , loss of control follows unless measures are taken to prevent it.

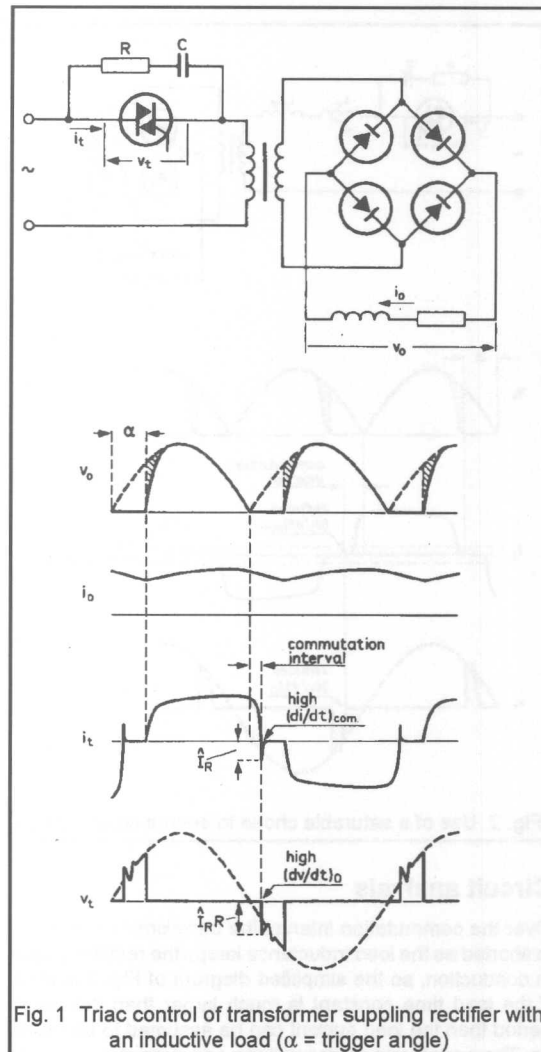
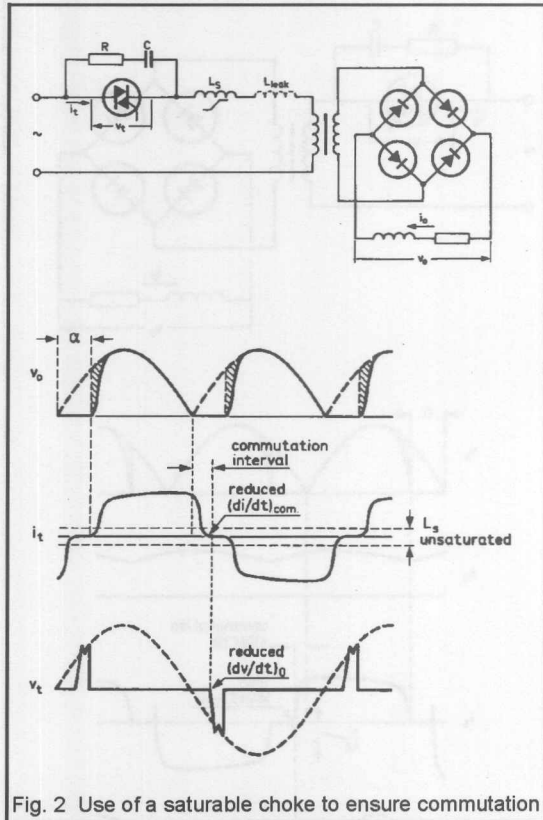


Fig. 1 Triac control of transformer supplying rectifier with an inductive load ( $\alpha$  = trigger angle)

### Obtaining reliable commutation

A saturable choke in series with the transformer primary proves effective in achieving reliable commutation (Fig. 2). Saturation should occur at a fraction of the rated load current so that the loss in the rectifier output voltage is minimised. At low currents the total inductance is large, thus softening the commutation and eliminating transients. The choke delays the rise in voltage so a quiescent period of a few tens of microseconds is introduced, during which time

the triac can recover. There is usually no difficulty in designing a choke such that the decay rate of current ( $di_{com}/dt$ ) and the rate of rise of voltage ( $dv/dt$ ) are sufficiently reduced to ensure reliable control.



### Circuit analysis

Over the commutation interval the transformer secondary is shorted as the load inductance keeps the rectifier diodes in conduction, so the simplified diagram of Fig. 3 applies. If the load time constant is much larger than the mains period then the load current can be assumed to be purely dc. The waveforms of triac voltage and current are given in Fig. 4. The mains voltage is given by  $v_i = V_s \sin \omega t$ . As the commutation interval is a fraction of the ac period then the rate of change of voltage during the commutation interval can be assumed to be linear, giving:

$$v_i = -\hat{V} \omega t \quad (1)$$

Over the period 0 to  $t_2$  the voltage across the saturable choke  $L_s$  and leakage inductance  $L_{leak}$  is equal to  $v_i$  (assuming the triac on-state voltage to be negligible). Assuming for this analysis that  $L_s$  remains in saturation (dashed portion in  $i_t$  waveform) then if  $L_{sat}$  is the saturated inductance, the following expression can be derived:

$$(L_{leak} + L_{sat}) \cdot di/dt = -\hat{V} \omega t \quad (2)$$

where  $di/dt$  is the rate of change of triac current.

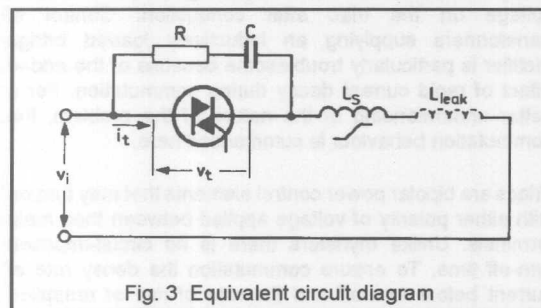


Fig. 3 Equivalent circuit diagram

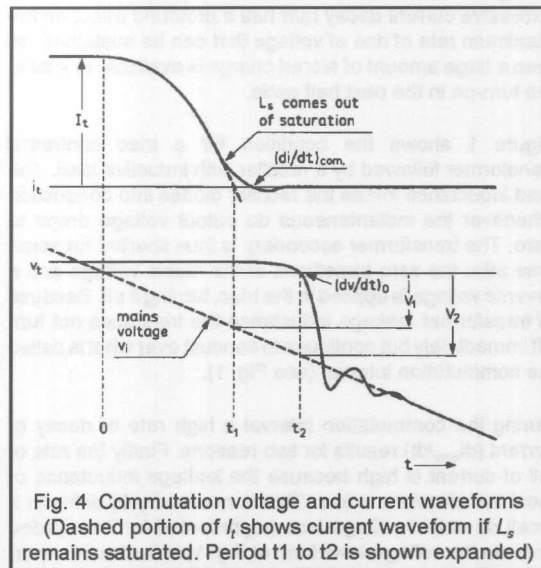


Fig. 4 Commutation voltage and current waveforms (Dashed portion of  $i_t$  shows current waveform if  $L_s$  remains saturated. Period  $t_1$  to  $t_2$  is shown expanded)

Integrating equation (2) gives:

$$i_t = I_t - \frac{\hat{V} \omega t^2}{2(L_{leak} + L_{sat})} \quad (3)$$

where  $I_t$  is the current prior to commutation.



At time  $t_1$ , current  $i_t$  passes through zero, so, from (3):

$$t_1 = \sqrt{\frac{2I_t(L_{leak} + L_{sat})}{\hat{V}\omega}} \quad (4)$$

At  $t_1$  the mains voltage has attained the value  $V_1$  which is found by combining equations (1) and (4) to give:

$$V_1 = -\sqrt{2\omega\hat{V}I_t(L_{leak} + L_{sat})} \quad (5)$$

Choke  $L_s$  comes out of saturation at low current levels so the triac turn-off point is delayed to time  $t_2$ . Since in a practical circuit the delay is only of the order of 50 $\mu$ s, the mains voltage  $V_2$  at the instant of turn-off is very nearly equal to  $V_1$ . Thus from equation 5:

$$V_2 \approx -\sqrt{2\omega\hat{V}I_t(L_{leak} + L_{sat})} \quad (6)$$

The triac conducts until time  $t_2$ . Denoting the value of unsaturated inductance as  $L_{unsat}$ , the current decay rate at zero current is given by:

$$\begin{aligned} \frac{di_c}{dt} &= \frac{V_2}{(L_{leak} + L_{unsat})} \\ &= -\frac{\sqrt{2\omega\hat{V}I_t(L_{leak} + L_{sat})}}{L_{leak} + L_{unsat}} \end{aligned} \quad (7)$$

The initial rate of rise of off-state voltage,  $dv_{com}/dt$ , can now be derived. This parameter is decisive for the behaviour of the triac, since a much greater  $dv/dt$  can be sustained after carrier recombination, that is, when the off-state voltage has reached a substantial value.

At time  $t_2$  the triac turns off but the voltage across it is still zero. The voltage drop across  $L_s$  and  $L_{leak}$  is equal to  $V_2$  and the rate of rise of current carried by these inductances,  $di_t/dt$ , is given in equation (7). The rate of rise of triac voltage  $dv/dt$  is determined by  $di_t/dt$  and the values of the snubber components R and C.

$$\frac{dv}{dt} = R \cdot \frac{di_t}{dt} + \frac{i}{C} \quad (8)$$

When the interval  $t_1$  to  $t_2$  is long enough, the triac has fully recovered at time  $t_2$ , and so the current  $i$  to be taken over by the parallel RC snubber network is zero. At time  $t_2$ ,  $dv/dt$  is equal to the initial rate of rise of voltage  $dv_0/dt$ . From equations (7) and (8):

$$\frac{dv_0}{dt} = \frac{R}{L_{leak} + L_{unsat}} \sqrt{2\omega\hat{V}I_t(L_{leak} + L_{sat})} \quad (9)$$

In circuits where no transformer is interposed between the triac and rectifier, some series inductance is still needed to restrict turn-on  $di/dt$ . In that case Equations (7) and (9) are still valid by omitting  $L_{leak}$ .

### Example - DC motor load

The motor control circuit of Fig. 5 illustrates the use of the design method proposed in the previous section. Since the motor has a fairly high inductance it may be considered as a constant current source, giving a severe test condition for triac commutation.

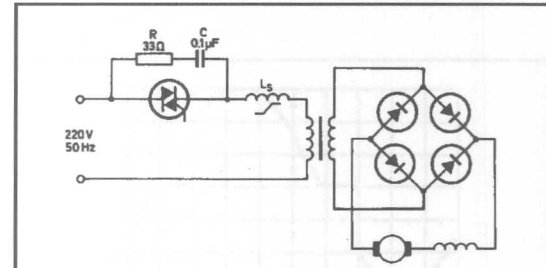


Fig. 5 DC Motor test circuit.

Choke	$L_{unsat} = 2.25\text{mH}$ , 30 turns on 36x23x10mm <sup>3</sup> toroid core
Transformer	220V/150V, 6kVA, 0.9mH leakage inductance
Motor	Series wound DC motor, Leakage inductance = 30mH

With  $L_{leak} = 0.9\text{mH}$ ,  $L_{unsat} = 2.25\text{mH}$  and  $L_{sat} \ll L_{leak}$  the circuit conditions can be calculated for a triac current of  $I_t = 20\text{A}$  and a 220V, 50Hz supply. Using equations (7) and (9) gives  $di_c/dt = -18.3\text{A/ms}$  and  $dv_0/dt = 0.6\text{V}/\mu\text{s}$ . These values can be compared with the commutation limits of the device to ensure that reliable commutation can be expected.

The inductance in the ac circuit also restricts turn-on  $di/dt$  which, for a continuous dc load current is:

$$\frac{di_{on}}{dt} \approx \frac{v_i}{t_{on}R} + \frac{v_i}{L_{leak} + L_{unsat}} \quad (10)$$

where  $v_i$  is the instantaneous ac input voltage,  $t_{on}$  is the turn-on time of the triac and R is the snubber resistance. Maximum turn-on  $di/dt$  occurs at the peak value of input voltage,  $v_i$ . The initial rise of on-state current depends on the snubber discharge current through R as well as the limiting effect of the circuit inductance.

The oscillograms of Figs. 6 to 10 illustrate circuit performance. With no choke added a large  $dv/dt$  was observed (Figs. 6 and 7) and so consequently commutation failed when motor current was increased to around 9A. As seen from Figs. 8 to 10 the choke softens commutation so that dependable control results even at 23A motor current. At this current (Fig. 10) the quiescent interval is about  $30\mu s$ , which is adequate time for the triac to recover.

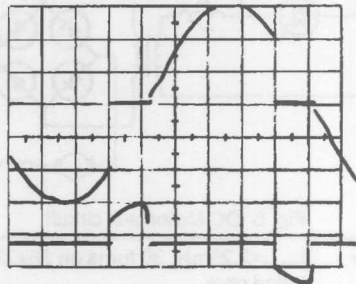


Fig. 6 Triac voltage and current. No series choke.  
7A motor current. Timebase: 2ms/div  
Upper trace: Triac voltage,  $v_t$  (100V/div)  
Lower trace: Triac current,  $i_t$  (5A/div)

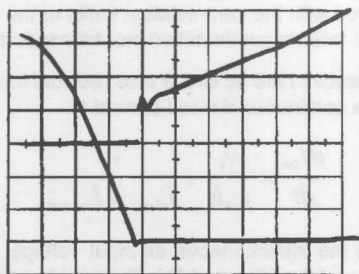


Fig. 7 Triac voltage and current. No series choke.  
7A motor current. N.B. Snap-off current  
Timebase: 100 $\mu s$ /div  
Upper trace: Triac voltage,  $v_t$  (20V/div)  
Lower trace: Triac current,  $i_t$  (1A/div)

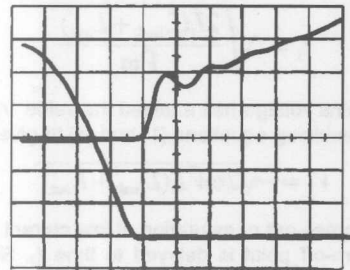


Fig. 8 Triac voltage and current. Series choke added.  
7A motor current. Timebase: 100 $\mu s$ /div  
Upper trace: Triac voltage,  $v_t$  (20V/div)  
Lower trace: Triac current,  $i_t$  (1A/div)

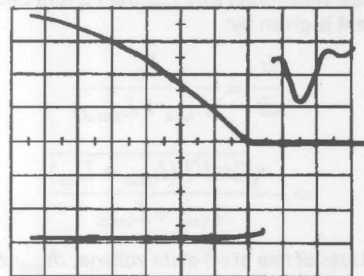


Fig. 9 Triac voltage and current. Series choke added.  
23A motor current. Timebase: 100 $\mu s$ /div  
Upper trace: Triac voltage,  $v_t$  (10V/div)  
Lower trace: Triac current,  $i_t$  (5A/div)

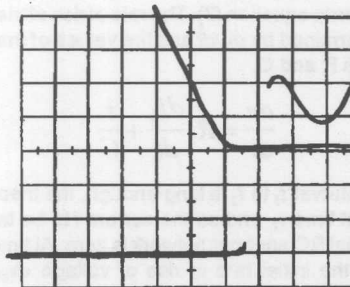


Fig. 10 Triac voltage and current. Series choke added  
23A motor current. N.B. slight reverse recovery current  
Timebase: 50 $\mu s$ /div  
Upper trace: Triac voltage,  $v_t$  (10V/div)  
Lower trace: Triac current,  $i_t$  (1A/div)

## 6.2.2 Domestic Power Control with Triacs and Thyristors

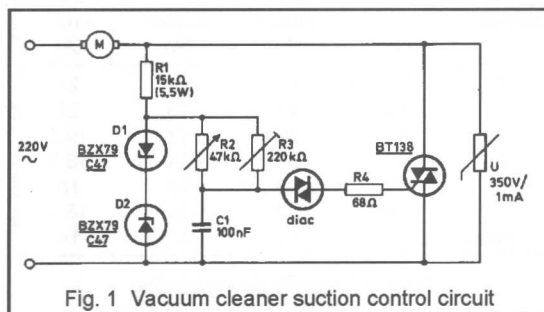
The increasing demand for more sophisticated domestic products can, in part, be met by providing the user with some form of electronic power control. This control can be used, for example, to adjust the suction of a vacuum cleaner, the brightness of room lighting or the speed of food mixers and electric drills.

It might be assumed that the cost of the electronics would be high, but this is not necessarily the case. With triacs and thyristors it is possible to produce high performance mains controllers which use only a few simple components. The following notes give details of some typical control circuits and highlight areas for special attention when adapting the designs for specific applications.

### Vacuum cleaner suction control

The competitive nature of the vacuum cleaner market has led to the development of a wide variety of machine types and accessories. In many cases, the speed of the motor remains constant and, if suction control is attempted, it consists merely of an adjustable vent in the air flow path. Electronic suction control sounds somewhat expensive and unnecessarily complicated for such an elementary application. In fact, by using a BT138 triac, a simple but nevertheless effective and reliable suction control circuit (Fig. 1) can be constructed very economically, and is suitable for all types of cleaner with a power consumption of up to 900W.

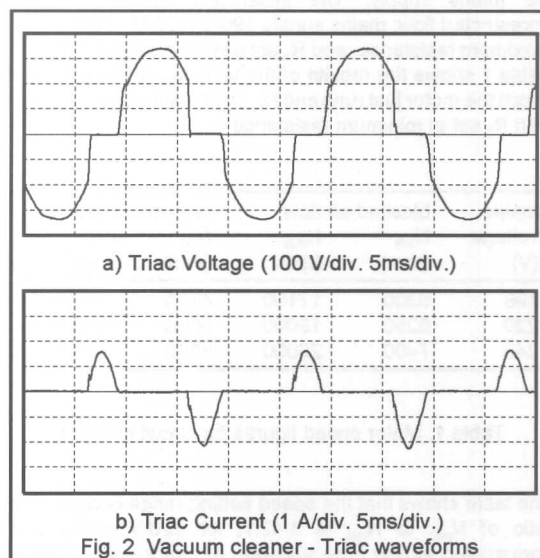
The heart of the circuit is the BT138. This is a glass passivated triac which can withstand high voltage bidirectional transients and has a very high thermal cycling performance. Furthermore its very low thermal impedance minimizes heatsink requirements.



### Circuit Description

In Fig. 1 the BT138 is the power control element. Its action is controlled by a diac which is switched on by a charge on  $C_1$  under the control of potentiometer  $R_2$ . The resistance of the diac is virtually infinite as long as the voltage across it

remains within the breakover voltage limits,  $-V_{BO}$  to  $+V_{BO}$ . During each half cycle of the mains sinewave,  $C_1$  charges until the voltage across it exceeds the diac breakover voltage. The diac then switches on and  $C_1$  discharges itself into the gate of the triac and switches it on. Diodes  $D_1$  and  $D_2$  stabilise the supply voltage to the charging circuit so that its operation is independent of mains voltage fluctuations. If  $-V_{BO}$  and  $+V_{BO}$  are equal and opposite, the triac will be triggered at the same time after the start of either a positive or negative half cycle. The conduction angle, and therefore the speed of the motor and the cleaner suction, is determined by the adjustment of  $R_2$ . Preset potentiometer  $R_3$  is used to set the minimum suction level. The width and amplitude of the trigger pulses are kept constant by gate resistor  $R_4$ . The zinc oxide voltage dependent resistor (U) minimises the possibility of damage to the triac due to very high voltage transients that may be superimposed on the mains supply voltage. Figure 2 shows the current and voltage waveforms for the triac when the conduction angle is  $30^\circ$ .



### Circuit Performance

A laboratory model of the circuit has been tested to determine the range of control that it has over the suction power of a typical vacuum cleaner. For the test, the cleaner was loaded with a water column. The result of the test is shown graphically in Fig. 3. The measured range of water column height (100 to 1100 mm) translates into a wide air flow range - from little more than a whisper to full suction.

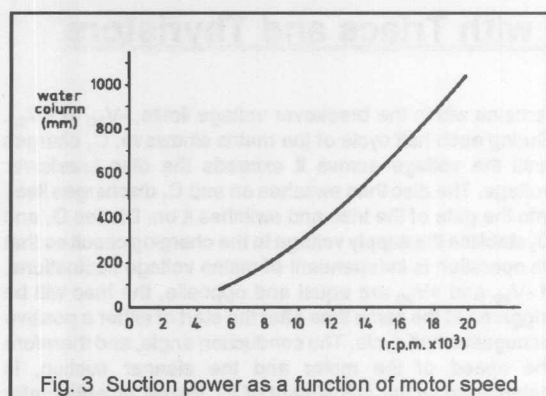


Fig. 3 Suction power as a function of motor speed

As suction power is a function of the speed of the vacuum cleaner motor, a second test was carried out to determine the range of motor speed control under conditions of minimum and maximum air flow (i.e. with the suction blocked and unrestricted). This test also checked the motor speed variation due to  $\pm 10\%$  variation of a nominal 220 V AC mains supply. The initial test conditions were: unrestricted flow; mains supply 198 V (220 V - 10%);  $R_2$  at maximum resistance, and  $R_3$  set so that the motor just ran. Table 1 shows the results of the test.  $N_{min}$  is the speed at which the motor just runs and  $N_{max}$  is the speed of the motor with  $R_2$  set at minimum resistance.

mains voltage (V)	blocked air flow		unrestricted air flow	
	$N_{min}$ (rpm)	$N_{max}$ (rpm)	$N_{min}$ (rpm)	$N_{max}$ (rpm)
198	5300	17100	4300	15400
220	6250	19000	5000	17100
242	7400	20000	6000	18200

Table 1. Motor speed figures for circuit of Fig. 1

The table shows that the speed setting range is wide. The ratio of  $N_{max}$  to  $N_{min}$  is 3.42:1, for 220 V mains and unrestricted airflow. The variation of motor speed due to variation of the mains input is quite small and represents a negligible change of suction. If  $D_1$  and  $D_2$  are omitted from the circuit, the speed setting ratio is reduced to 1.82:1 under the same conditions. The table also shows that the difference between the  $N_{min}$  for minimum and maximum air flow is quite small. This implies that speed stabilisation is unnecessary.

### Special Design Considerations

The circuit shown in Fig. 1 has been shown to work well in a typical vacuum cleaner application. But motors and environments do vary, so some aspects of the design should be looked at carefully before it is finalised.

#### Circuit positioning

The siting of the circuit, within the case of the cleaner, is particularly important. In some areas within the cleaner the temperature can be quite high. The circuit, and in particular the triac and its heatsink, should not be placed in one of these areas if the designer is to avoid problems keeping the temperature of the triac below  $T_{jmax}$ .

#### Starting current

Another factor that may lead to thermal problems is that of inrush current. The starting current of a vacuum cleaner motor is typically as shown in Fig. 4. The rms current during the first 20 ms could be 20 A or more. The current decays to its steady state value in about 1 s. To ensure that the triac does not overheat, reference should be made to the inrush current curves in the triac data sheet, the curve for the BT138 is reproduced in Fig. 5.

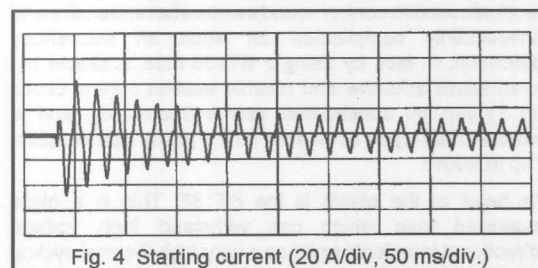


Fig. 4 Starting current (20 A/div, 50 ms/div.)

cycle number.	time (ms)	peak current (A)	rms current (A)	'limit' current (A)
1	20	49	22	24
2	40	41	18	21
3	60	35	13	19.5
4	80	32	14	18.5
5	100	29	13	18
10	200	20	9	15.5
20	400	14	6.3	14

Table 2. Currents during starting

The first step in checking for a problem is to estimate the mounting base temperature,  $T_{mb}$ , prior to starting. A reasonable figure would be the worst case steady state value of  $T_{mb}$  during normal running.



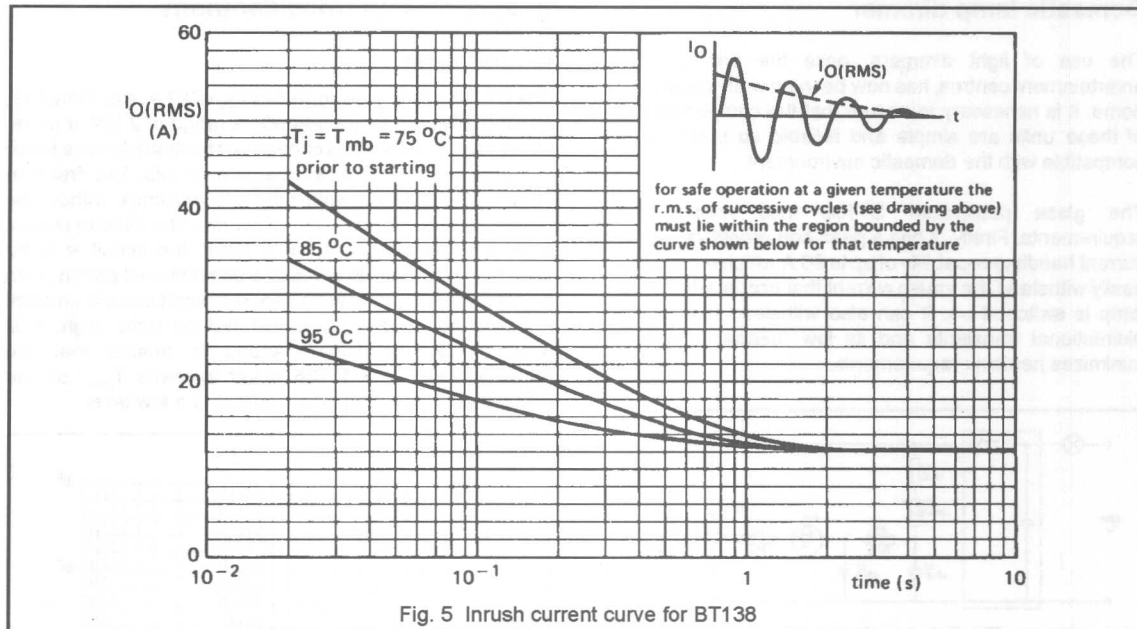


Fig. 5 Inrush current curve for BT138

Step 2 is to calculate the rms value of one cycle of the starting current at several times during start up and step 3 is to compare these figures with the values taken from the appropriate line of the inrush current curve.

As an example consider the performance of the BT138 driving a motor whose starting current is shown in Fig. 4. Direct measurement indicated that during normal running the  $T_{mb}$  of a BT138 mounted on a particular heatsink, would be no more than 22°C above ambient. From other measurements it was estimated that the ambient temperature would not exceed 73°C. These figures give a worst case steady state  $T_{mb}$  of 95°C. It can be assumed that this is the highest temperature that the mounting base could be, prior to starting - a reasonable assumption which covers the case where the motor has been running for a long time, is turned off and then started again before there has been any cooling.

The rms values of cycles 1 to 5, 10 and 20 of the starting current are given in Table 2. Since the current is not an ideal sine wave these have been calculated from the peak current by assuming a crest factor (peak to rms) of 2.23. Also shown are the relevant  $I_{O(RMS)}$  figures from the 95°C line of Fig. 5. Since 'actual' inrush current is always less than the 'allowed' current it is safe to use the BT138 under the proposed conditions to control the motor. It should be noted that because the crest factor is  $>\sqrt{2}$  the dissipation of the BT138 will be less than assumed by the inrush current curves of Fig. 5.

## Commutation

The circuit shown in Fig. 1 has no RC snubber. This was because the values of  $di/dt$  and  $dV/dt$  generated by the circuit were well within the capability of the BT138. This will often be the case with vacuum cleaner motors for two reasons:

- these motors introduce only a small phase shift in the current, so the voltage step is small and the  $dV/dt$  is low,
- the steady state value of the current is much less than the maximum rating of the BT138, this amounts to a  $di/dt$  well within the capability of the BT138.

However care must be taken to ensure that this is true in all applications. In particular, care should be taken to ensure that the triac switches correctly even during starting. If a snubber is found to be necessary then a 100  $\Omega$  0.5 W resistor in series with a 0.1  $\mu F$  capacitor will be more than adequate in most circumstances.

## Interference

It is, of course, necessary to check that the overall equipment complies with local regulations for conducted and radiated interference. However, the measures taken to suppress the electrical 'noise' of the motor combined with the motor itself will often be more than sufficient to overcome the interference generated by the switching of the triac but this must be checked in all applications.

## Domestic lamp dimmer

The use of light dimmers, once the prerogative of entertainment centres, has now become widespread in the home. It is necessary to ensure that the component parts of these units are simple and reliable so that they are compatible with the domestic environment.

The glass passivated BT138 triac meets these requirements. Firstly, it has a peak non-repetitive on-state current handling capability of up to 90 A which means it can easily withstand the inrush current that occurs when a cold lamp is switched on. It can also withstand high voltage bidirectional transients and its low thermal impedance minimizes heatsink requirements.

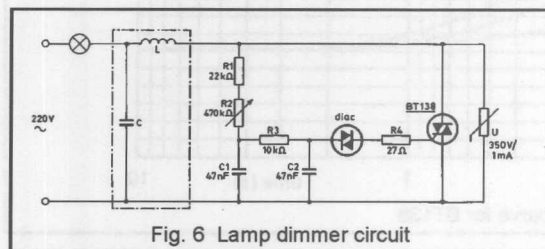


Fig. 6 Lamp dimmer circuit

## Circuit Description

A simple circuit of a light dimmer using the BT138 is given in Fig. 6. The BT138 is the power control element, triggered via the diac. The setting of potentiometer  $R_2$  determines the phase difference between the mains sine wave and the voltage across  $C_2$ . This in turn sets the triac triggering angle and the lamp intensity.

The resistance of the diac is very high as long as the voltage across it remains within its breakover voltage limits,  $-V_{BO}$  to  $+V_{BO}$ . Each half cycle of the mains charges  $C_2$  via  $R_1$ ,  $R_2$  and  $R_3$  until the voltage being applied to the diac reaches one of its breakover levels. The diac then conducts and  $C_2$  discharges into the gate of the triac, switching it on. If  $-V_{BO}$  and  $+V_{BO}$  are equal and opposite, the triac will be triggered at the same time after the start of either a positive or negative half cycle. If  $C_1$  were not included in the circuit, the voltage across  $C_2$  would change abruptly after triggering and cause the phase relationship between the mains voltage and voltage across  $C_2$  to progressively alter. This would cause an undesirable hysteresis effect. The voltage across  $C_1$  partially restores the voltage across  $C_2$  after triggering and thereby minimizes the hysteresis effect. The width and amplitude of the trigger pulses are kept constant by gate resistor  $R_4$ . The VDR minimizes the possibility of the triac being damaged by high voltage transients that may be superimposed on the mains supply voltage.

## Special Design Considerations

### Circuit rating

The BT138 has an rms current rating of 12 A. It is, therefore, capable of controlling loads with a rating of 2 kW or more. However, the load of this circuit must be restricted to a much lower level. There are two reasons for this. The first is to keep mains distortion within the allowed limits, without the necessity of expensive filter networks. The second reason is to limit dissipation. If, as is likely, the circuit is to be mounted in the wall in place of a conventional switch, then air circulation is going to be very restricted and the ambient temperature around the circuit will be quite high. It is important for reliability reasons to ensure that the temperature of the BT138 never exceeds  $T_{jmax}$ , so the dissipation of the triac must be kept to a low level.

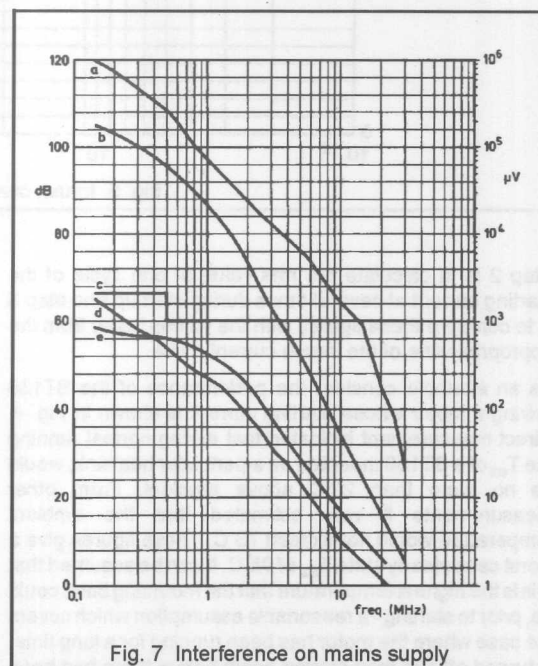


Fig. 7 Interference on mains supply

## Interference

Regulations concerning conducted and radiated interference vary considerably from country to country but it is likely that some form of filter will be needed. The simple LC filter shown within the dashed-lined box in Fig. 6 is often all that is needed. The values of the filter components will vary, but a combination of 0.15  $\mu$ F capacitor and a low Q inductor of 2.5  $\mu$ H was found to be sufficient for the circuit to meet the C.I.S.P.R. limits. This is illustrated by the plots shown in Fig. 7. Curves (a) and (b) show the level of noise on the mains supply for the circuit, without filter, when

controlling 550 W and 25 W loads respectively. Curves (d) and (e) are for the circuit with filter connected showing that the C.I.S.P.R. limit, which is curve (c), has been met.

### Filter inductor

Having selected the value of filter inductor, the designer has then to decide how to make it. Construction will not be too critical - it is not necessary to achieve a high Q - and there will be considerable room for reducing its size. However, care must be taken to ensure that the inductor does not saturate when the inrush current of a cold lamp flows through it. If the inductor does saturate then the filter capacitor will, effectively, be shorted out by the triac. In this case the triac current could rise faster than the  $di/dt$  rating allows. This could cause progressive damage to the triac resulting in premature failure.

### Speed control for food mixers and electric drills

Food mixers and electric hand drills are products whose useability is improved by the addition of electronic speed control. But they are products where costs have to be tightly controlled so the choice of circuit is very important. This decision is made harder by the need to have a good speed regulation under the widely varying loads that these products are subjected to.

The circuits to be described provide continuous control of motor speed over a wide speed range by adjusting the conduction angle of a BT151 thyristor. They compensate for load variation by adjusting the firing angle when there is a change in the motor speed - as indicated by a change in its back EMF.

### Back EMF Feedback Circuits

A simple motor speed control circuit that employs back EMF to compensate for changes in motor load and mains voltage is shown in Fig. 8(a). The resistor chain  $R_1$ ,  $R_2$ ,  $R_3$  and diode  $D_1$  provide a positive going reference potential to the thyristor gate via diode  $D_2$ . Diode  $D_1$  is used to reduce the dissipation in the resistor chain by some 50% and diode  $D_2$  isolates the trigger circuit with the thyristor in the on-state. When the thyristor is not conducting the motor produces a back EMF voltage across the armature proportional to residual flux and motor speed. This appears as a positive potential at the thyristor cathode.

A thyristor fires when its gate potential is greater than cathode potential by some fixed amount. Depending on the waveform shape and amplitude at the gate, the circuit may function in several modes.

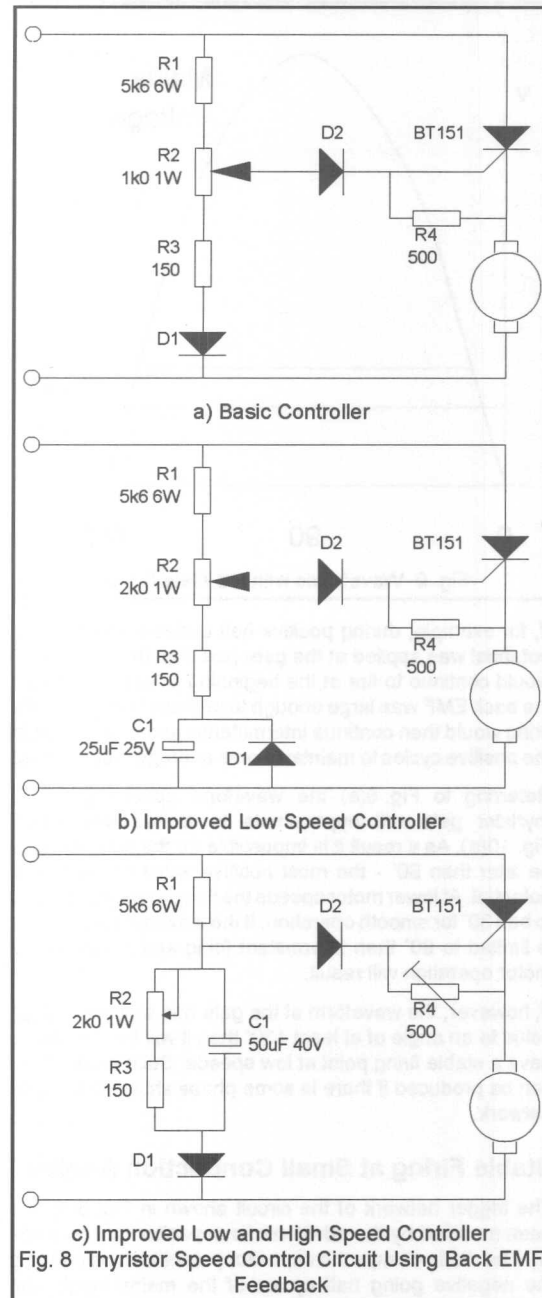
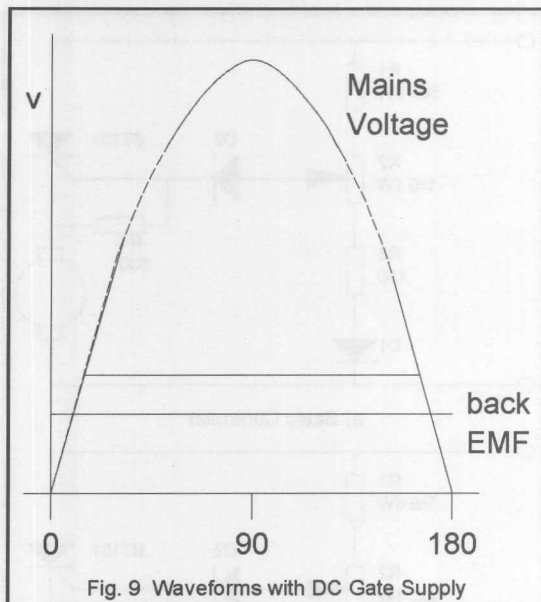


Fig. 8 Thyristor Speed Control Circuit Using Back EMF Feedback



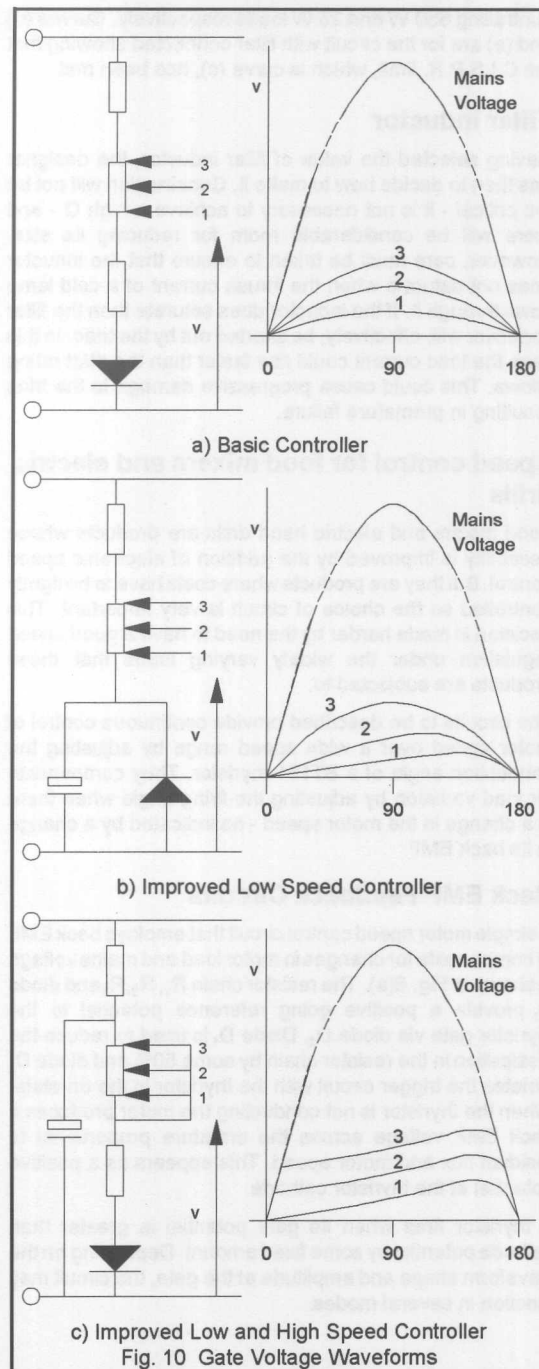
If, for example, during positive half cycles a constant DC potential was applied at the gate (see Fig. 9), the thyristor would continue to fire at the beginning of each cycle until the back EMF was large enough to prevent firing. Thyristor firing would then continue intermittently at the beginning of the positive cycles to maintain some average motor speed.

Referring to Fig. 8(a) the waveform appearing at the thyristor gate will approximate to a half sine wave, Fig. 10(a). As a result it is impossible for the firing angle to be later than  $90^\circ$  - the most positive value of the trigger potential. At lower motor speeds the firing angle might need to be  $130^\circ$  for smooth operation. If the maximum firing angle is limited to  $90^\circ$  then intermittent firing and roughness of motor operation will result.

If, however, the waveform at the gate has a positive slope value to an angle of at least  $130^\circ$  then it will be possible to have a stable firing point at low speeds. Such a waveform can be produced if there is some phase shift in the trigger network.

### Stable Firing at Small Conduction Angles

The trigger network of the circuit shown in Fig. 8(b) has been modified by the addition of a capacitor  $C_1$  and diode  $D_1$ . The diode clamps the capacitor potential at zero during the negative going half cycles of the mains input. The waveform developed across the capacitor has a positive slope to some  $140^\circ$ , allowing thyristor triggering to be delayed to this point.





As the slider of  $R_2$  is moved towards  $R_1$ , the peak of the waveform at the gate will move towards  $90^\circ$  as shown in Fig. 10(b). As the speed increases, the no load firing angle will also advance by a similar amount so stability will be maintained. This circuit will give smoother and more stable performance than the circuit of Fig. 8(a). It will, however, give a marginally greater speed drop for a given motor loading at low speed settings. At the maximum speed settings the circuit of Fig. 8(a) approximates to that of Fig. 8(b).

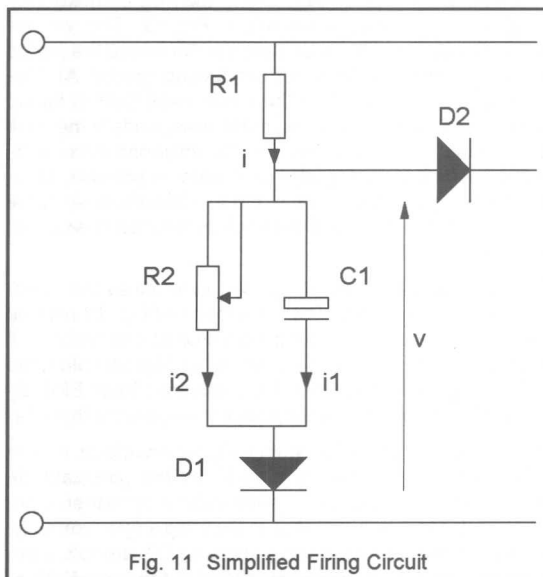


Fig. 11 Simplified Firing Circuit

### Improved Motor Performance With Stable Firing

Both the circuits so far discussed have gate voltage waveforms that are of near linear slope from the zero point of each positive half cycle, see Figs. 10(a) and (b). This means that the only time that the thyristor can be fired early in the mains cycle, say at  $20^\circ$ , is when the back EMF and hence motor speed is very low. This effect tends to prevent smooth running at high speeds and high loads.

Stable triggering, at low angles, can be achieved if the gate voltage ramp starts each cycle at a small positive level. This means that the time to reach the minimum trigger voltage is reduced. The circuit of Fig. 8(c) is one way of achieving this. In this circuit capacitor  $C_1$  is charged during positive half cycles via resistor  $R_1$  and diode  $D_1$ . During negative half cycles the only discharge path for capacitor  $C_1$  is via resistors  $R_2$  and  $R_3$ .

Diode  $D_1$  also prevents  $C_1$  from being discharged as the thyristor switches off by the inductively generated pulse from the motor. As the value of resistor  $R_2$  is increased, capacitor  $C_1$  is discharged less during negative half cycles but its charging waveform remains substantially unchanged. Hence the result of varying  $R_2$  is to shift the DC level of the ramp waveform produced across  $C_1$ .

Diode  $D_2$  isolates the triggering circuit when the thyristor is ON. Resistor  $R_4$  adjusts minimum speed, and by effectively bleeding a constant current, in conjunction with the gate current from the triggering circuit, it enables resistor  $R_2$  to give consistent speed settings.

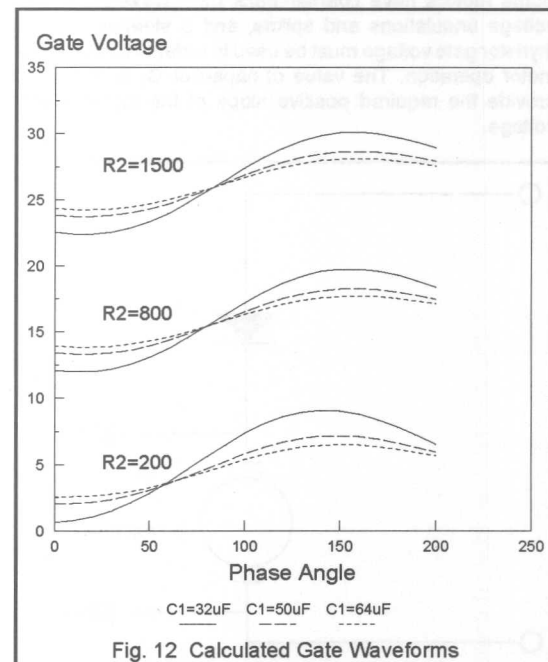


Fig. 12 Calculated Gate Waveforms

### Circuit Design

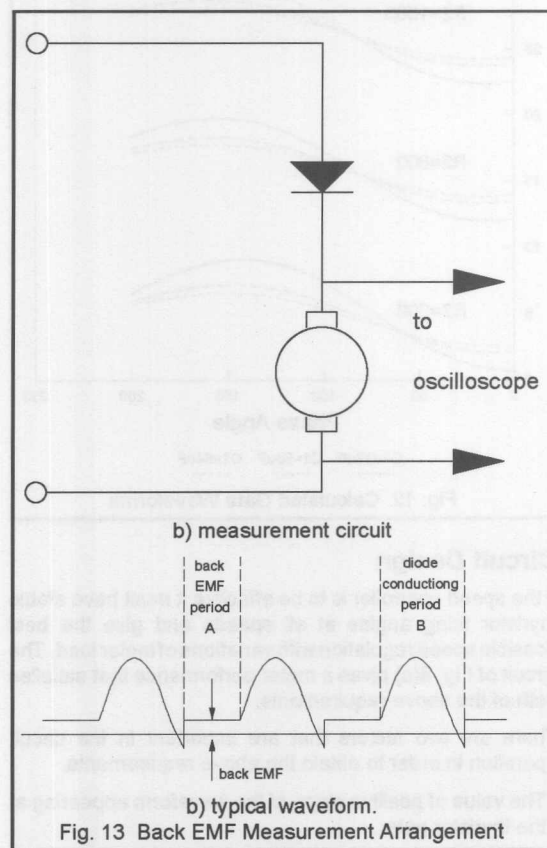
If the speed controller is to be effective it must have stable thyristor firing angles at all speeds and give the best possible speed regulation with variations of motor load. The circuit of Fig. 8(c) gives a motor performance that satisfies both of the above requirements.

There are two factors that are important in the circuit operation in order to obtain the above requirements.

- The value of positive slope of the waveform appearing at the thyristor gate.
- The phase angle at which the positive peak gate voltage is reached during a positive half cycle of mains input.

As previously described the charging of capacitor  $C_1$  by resistor  $R_1$  determines the rate of rise of voltage at the thyristor gate during the positive half cycle. However, resistor  $R_1$  must also have a value such that several times the maximum thyristor gate current passes through the RC network to  $D_1$ . This current will then give consistent speed settings with the spread of thyristor gate currents when the minimum speed is set by resistor  $R_4$ .

The positive slope value of the thyristor gate voltage will have to be fixed according to the motor used. A motor that gives a smooth back EMF voltage will allow a low slope value to be used, giving good torque speed characteristics. Some motors have coarser back EMF waveforms, with voltage undulations and spikes, and a steeper slope of thyristor gate voltage must be used in order to obtain stable motor operation. The value of capacitor  $C_1$  is chosen to provide the required positive slope of the thyristor gate voltage.



Some calculations have been made on the circuit of Fig. 8(c) simplified to the form of Fig. 11, where it is assumed that current flowing to the thyristor gate is small compared with the current flowing through resistor  $R_1$ . An expression has been derived for the voltage that would appear at the anode of  $D_2$  in terms of  $R_1$ ,  $R_2$  and  $C_1$  and is given later. Component values have been substituted into the expression to give the thyristor gate waveforms shown in Fig. 12.

In order to adjust the circuit to suit a given motor, the back EMF of the motor must be known. This may be measured using the arrangement shown in Fig. 13. The voltage appearing across the motor is measured during the period when the series diode is not conducting (period A). The voltage so obtained will be the motor back EMF at its top speed on half wave operation, and corresponds to the back EMF that would be obtained from the unloaded motor at its highest speed when thyristor controlled. In practice, since the mains input is a sine wave, there is little increase in the 'no load' speed when the firing angle is reduced to less than about  $70^\circ$ .

The value of resistor  $R_2$  in Fig. 8(c) determines the motor 'no load' speed setting. The waveforms of Fig. 12 may be used as a guide to obtaining the value of this resistor. It must be chosen so that at  $70^\circ$  and at its highest value, the gate voltage is higher than the measured back EMF by about 2 V - the forward gate/cathode voltage of the thyristor.

The thyristor is turned ON when a trigger waveform, shown in Fig. 12, exceeds the back EMF by the gate/cathode voltage. So, if the back EMF varies within a cycle then there will be a cycle to cycle variation in the firing angle. Normally, random variations of the firing angle by  $20^\circ$  are tolerable. If, for example, there were variations in the back EMF of 1 V, then with a firing angle of  $70^\circ$  and a capacitor of  $32 \mu\text{F}$ , the variation of firing angle would be about  $12^\circ$ . With capacitor values of  $50 \mu\text{F}$  and  $64 \mu\text{F}$  the firing angles variations would be  $19^\circ$  and  $25^\circ$  respectively. Therefore, a capacitor value of  $50 \mu\text{F}$  would be suitable.

## Performance

The torque speed characteristics of the three circuits, when used to drive an electric drill, are compared in Fig. 14. It may be seen that the circuit of Fig. 8(b) has a poorer performance than the two other circuits. That of Fig. 8(c) may be seen to give a similar performance to the circuit of Fig. 8(a) at low speeds but, at high speeds and torques, it is better. It should be noted that the circuits of Figs. 8(b) and (c) provide low speed operation free from the intermittent firing and noise of the Fig. 8(a) circuit. Figure 15 compares the circuits of Fig. 8(a) and 8(c) when the load is a food mixer motor.

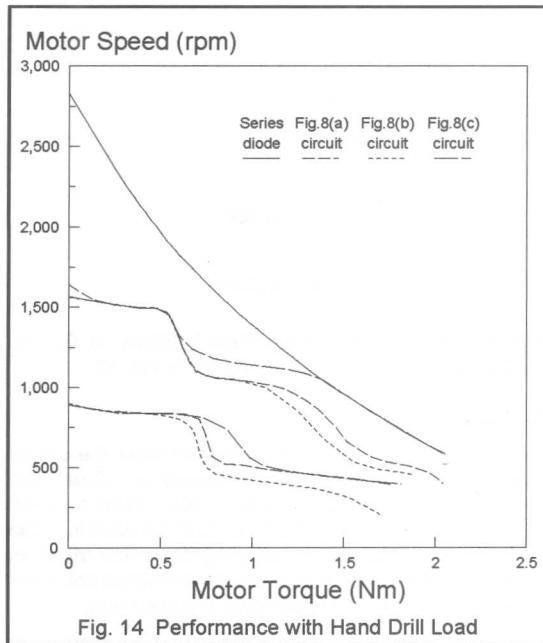


Fig. 14 Performance with Hand Drill Load

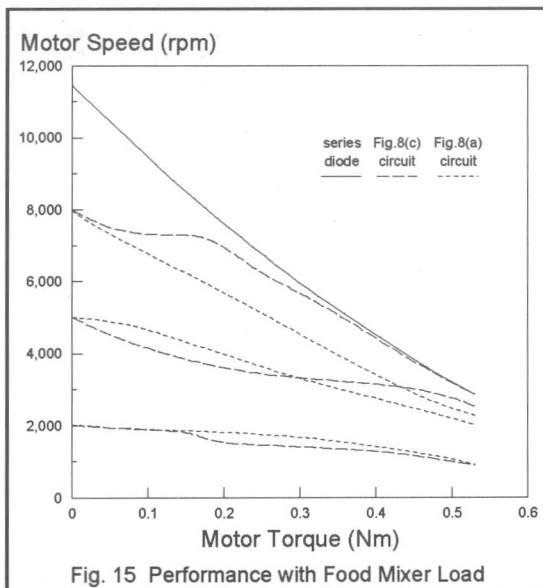


Fig. 15 Performance with Food Mixer Load

### Circuit Calculations

The following analysis derives an expression for voltage 'v' at the anode of  $D_2$ . This expression can be used to produce

the gate voltage waveforms shown in Fig. 12. The analysis assumes that the current drawn by the thyristor gate is negligible in comparison with the current flowing in  $R_1$ .

The charging current  $i_1$  for capacitor  $C_1$  in Fig. 11, is given by:

$$i_1 = \frac{dq}{dt} = C_1 \frac{dv}{dt}$$

and

$$i_2 = \frac{v}{R_2}$$

Representing a mains half sine wave by  $f(E)$  where  $E$  is the peak mains voltage.

$$i = \frac{f(E) - v}{R_1} = i_1 + i_2$$

therefore,

$$\frac{f(E) - v}{R_1} = C_1 \frac{dv}{dt} + \frac{v}{R_2}$$

where  $i$ ,  $i_1$ ,  $i_2$  are instantaneous currents.

Simplifying:-

$$C_1 \frac{dv}{dt} + v \left( \frac{1}{R_1} + \frac{1}{R_2} \right) = \frac{f(E)}{R_1}$$

Fourier analysis of a half sinewave gives:-

$$f(E) = E \left\{ \frac{1}{\pi} + \frac{1}{2} \sin(\theta) - \frac{2}{\pi} \sum_{n=2,4,6,\dots}^{\infty} \frac{\cos(n\theta)}{n^2 - 1} \right\}$$

neglecting terms of the Fourier series with  $n > 2$ , then

$$C_1 \frac{dv}{dt} + v \left( \frac{1}{R_1} + \frac{1}{R_2} \right) = \frac{E}{R_1} \left\{ \frac{1}{\pi} + \frac{1}{2} \sin(\omega t) - \frac{2}{3\pi} \cos(2\omega t) \right\}$$

then

$$C_1 \frac{dv}{dt} + v \left( \frac{1}{R_1} + \frac{1}{R_2} \right) - \frac{E}{R_1 \pi} = \frac{E}{R_1} \left\{ \frac{1}{2} \sin(\omega t) - \frac{2}{3\pi} \cos(2\omega t) \right\} \quad (1)$$

simplifies to

$$A \frac{dv}{dt} + Bv - D = X \sin(\omega t) - Y \cos(2\omega t) \quad (2)$$

where  $A$ ,  $B$ ,  $D$ ,  $X$ ,  $Y$  are constants.

Put

$$v = a \sin(\omega t) + b \cos(\omega t)$$

$$+ c \sin(2\omega t) + d \cos(2\omega t) + \frac{D}{B} \quad (3)$$

where  $a$ ,  $b$ ,  $c$ ,  $d$  are constants.

$$\frac{dv}{dt} = a\omega \cos(\omega t) - b\omega \sin(\omega t) + 2c\omega \cos(2\omega t) - 2d\omega \sin(2\omega t) \quad (4)$$

substituting (3) and (4) in equation (2) and equating terms in  $\cos(\omega t)$ ,  $\cos(2\omega t)$ ,  $\sin(\omega t)$ ,  $\sin(2\omega t)$ , then

$$v = \frac{BX}{A^2\omega^2 + B^2} \cdot \sin(\omega t) - \frac{A\omega X}{A^2\omega^2 + B^2} \cdot \cos(\omega t) - \frac{2A\omega Y}{4A^2\omega^2 + B^2} \cdot \sin(2\omega t) - \frac{BY}{4A^2\omega^2 + B^2} \cdot \cos(2\omega t) + \frac{D}{B}$$

substituting for the constants in equation (2) gives:

$$v = R_2 E \cdot \frac{(R_1 + R_2) \sin(\omega t) - \omega C_1 R_1 R_2 \cos(\omega t)}{2\omega^2 C_1^2 R_1^2 R_2^2 + (R_1 + R_2)^2} + R_2 E \cdot \frac{1}{\pi(R_1 + R_2)} - R_2 E \cdot \frac{4\omega C_1 R_1 R_2 \sin(2\omega t) + 4(R_1 + R_2) \cos(2\omega t)}{3\pi[4\omega^2 C_1^2 R_1^2 R_2^2 + (R_1 + R_2)^2]}$$

This may be simplified since

$$(R_1 + R_2)^2 \ll 2\omega^2 C_1^2 R_1^2 R_2^2$$

So the voltage that the trigger circuit would apply to the gate (assuming the gate draws no current) is given by:

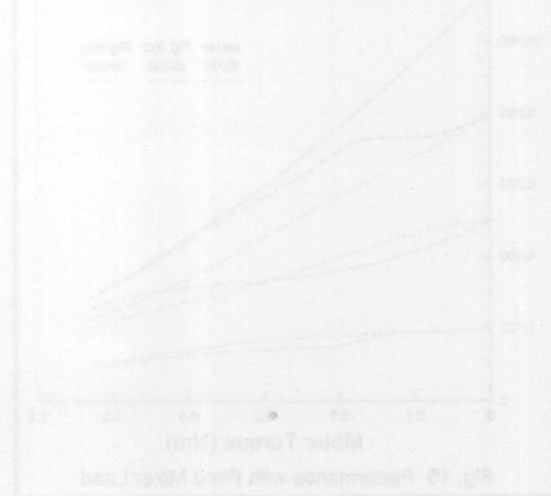
$$v = \frac{R_2 E}{\pi(R_1 + R_2)} + \frac{R_2 E}{2\omega^2 C_1^2 R_1^2 R_2^2} \{ (R_1 + R_2) \sin(\omega t) - \omega C_1 R_1 R_2 \cos(\omega t) - \frac{2}{3\pi} \omega C_1 R_1 R_2 \sin(2\omega t) - \frac{1}{3\pi} (R_1 + R_2) \cos(2\omega t) \}$$

Solving this equation for a different values of  $C_1$  and positions of  $R_2$  gives the curves shown in Fig. 12.

### Conclusions

The addition of electronic control can enhance the overall useability of many domestic products. Cost and performance requirements are major factors when determining the type of control circuit to be used in these applications. It is possible, using thyristors and triacs, to construct a range of phase control circuits which can meet many of these cost and operational requirements.

Although these circuits are not complex and use only simple components, it is still important to design with care to ensure that the best performance is achieved. This report has given examples of some of these circuits and has highlighted the areas of their design requiring particular care.





### 6.2.3 Design of a Time Proportional Temperature Controller

Electronic temperature control is no longer new: phase and on/off controls for heaters have been widely used to replace mechanical switches. However, both phase control and on/off control have disadvantages. Conventional phase control allows fully-proportional control of the power dissipated in the load, but the high rates of change of current and voltage cause RFI and transients on the mains supply. Because of this effect, phase control is not allowed to be used for domestic heaters. Simple on/off control with zero-voltage switching avoids generation of RFI but the amount of hysteresis required to prevent temperature oscillations does not give the required control accuracy.

#### The principle of time-proportional control

Time proportional control combines the zero-voltage switching of on/off control with the accuracy of proportional control and so eliminates the disadvantages of these two alternative systems. Time-proportional control regulates the load power such that there will be no overshoot or undershoot of the desired temperature as is the case with normal on/off systems. The TDA1023 has been designed to provide time-proportional control for room heaters and electric heating elements using a minimum number of external components. It incorporates additional features to provide fail-safe operation and fine control of the temperature.

There are three states of operation when using time-proportional control:

- load switched fully off,
- load power proportional to the difference between actual and desired temperatures,
- load switched fully on.

Figure 1 illustrates the principle; the load is switched on once and off once in a fixed repetition period, the ratio of the on and off periods providing the proportional control. This method of control can cause mains flicker; the mains voltage changes slightly each time the load is switched on or off.

CENELEC, the European Committee for Electro-technical Standardisation, has published rules which limit the rate at which domestic heating apparatus may be switched on and off. Table 1 gives the minimum repetition period for a range of load powers and common mains voltages from CENELEC publication EN50.006.

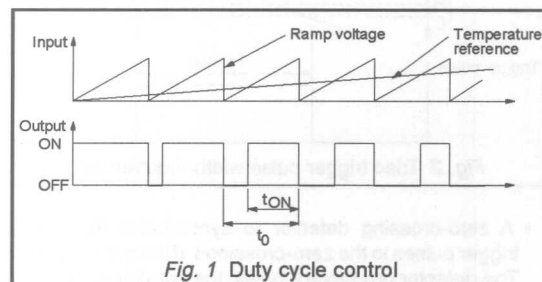


Fig. 1 Duty cycle control

Appliance Power (W)	Repetition period, $t_0$ (s)		
	220V	240V	380V
600	0.2	0.2	
800	0.8	0.3	0.1
1000	2.0	1.0	0.2
1200	4.6	2.0	0.2
1400	7.0	4.3	0.2
1600	10.0	6.3	0.3
1800	16.0	8.9	0.5
2000	24.0	13.0	0.9
2200	32.0	17.0	1.3
2400	40.0	24.0	1.9
2600		31.0	2.6
2800			3.6

Table 1. CENELEC minimum repetition periods for Domestic Heater Applications

#### Description of the TDA1023

The TDA1023 is a 16-pin dual in-line integrated circuit designed to provide time-proportional power control of electrical heating elements. The TDA1023 is ideally suited for the control of:

- Panel heaters
- Cooker elements
- Electric irons
- Water heaters
- Industrial applications, e.g. temperature controlled oil baths, air conditioners.

The TDA1023 incorporates the following functions:

- A stabilised power supply. The TDA1023 may be connected directly to the AC mains using either a dropping resistor or capacitor. It provides a stabilised reference voltage for the temperature-sensing network.

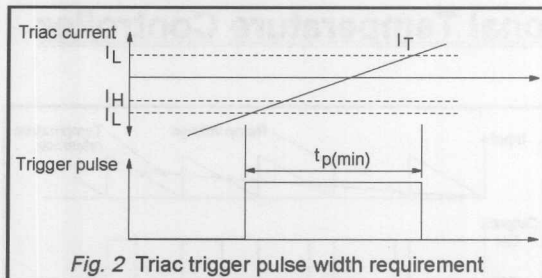


Fig. 2 Triac trigger pulse width requirement

- A zero-crossing detector to synchronise the output trigger pulses to the zero-crossings of the mains supply. The detector produces a pulse, the duration of which is determined by an external resistor, centred on the zero-crossing of the mains voltage.
- A comparator with adjustable hysteresis, preventing spurious triggering of the output. This compares a thermistor voltage, a function of the room temperature, with the voltage from the temperature selection dial.
- A voltage translation circuit for the potentiometer input. Normally, the relatively small temperature variation in a room ( $5^{\circ}\text{C}$  to  $30^{\circ}\text{C}$ ) corresponds to a narrow angle of rotation of a potentiometer shaft. Use of this circuit doubles the angle of rotation of the potentiometer shaft for the same temperature range.
- A sensor fail-safe circuit to prevent triggering if the thermistor input becomes open or short-circuited.

• A timing generator with an adjustable proportional band. This allows a full 100% control of the load current over a temperature range of only  $1^{\circ}\text{C}$  or  $5^{\circ}\text{C}$ . The repetition period of the timing generator may be set by an external capacitor to conform to the CENELEC specifications for mains load switching.

• An output amplifier with a current-limited output. The amplifier has an output current capability of at least 200mA and is stabilised to 10V while the current limit is not exceeded.

• Input buffers, to isolate the voltage translation circuit and comparator from external influences.

• A control gate circuit to activate the output if there is a mains zero-crossing, the comparator is ON and the fail-safe comparator is OFF.

Although designed specifically for time proportional control, the TDA1023 is also suitable for applications requiring on/off control if the timing generator is not used.

### Required Duration of Triac Trigger Pulse

The main advantage of triggering at the instant when the applied voltage passes through zero is that this mode of operation renders the use of RF suppression components unnecessary. For time-proportional control, continuous conduction of the triac may be required for many cycles of the mains supply. To maintain conduction while the load current is approaching the zero-crossing, the trigger pulse must last from the time when the load current falls to the value of the triac holding current ( $I_H$ ), until the time when the load current reaches the triac latching current ( $I_L$ ).

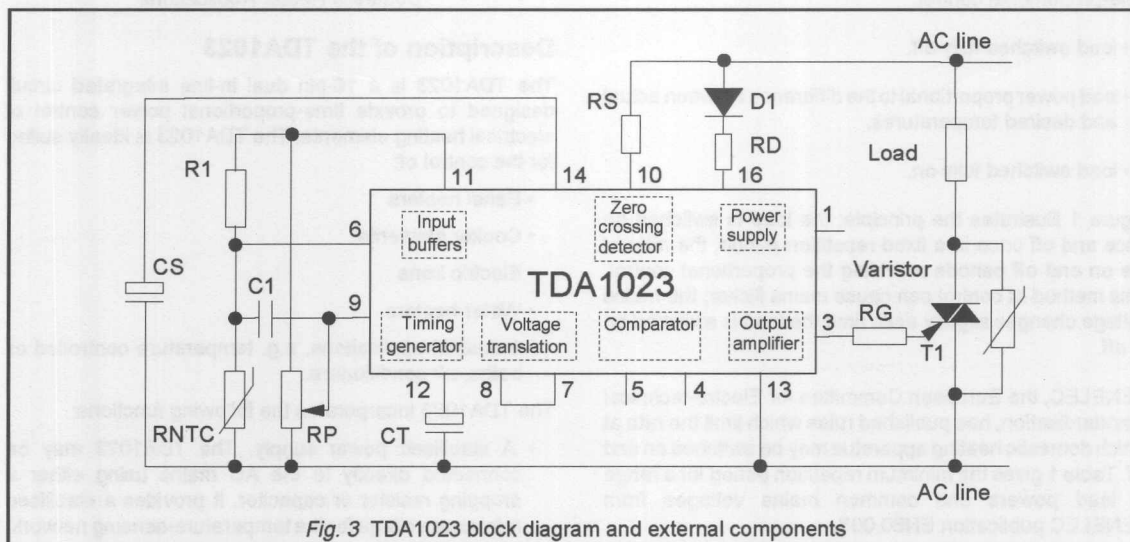
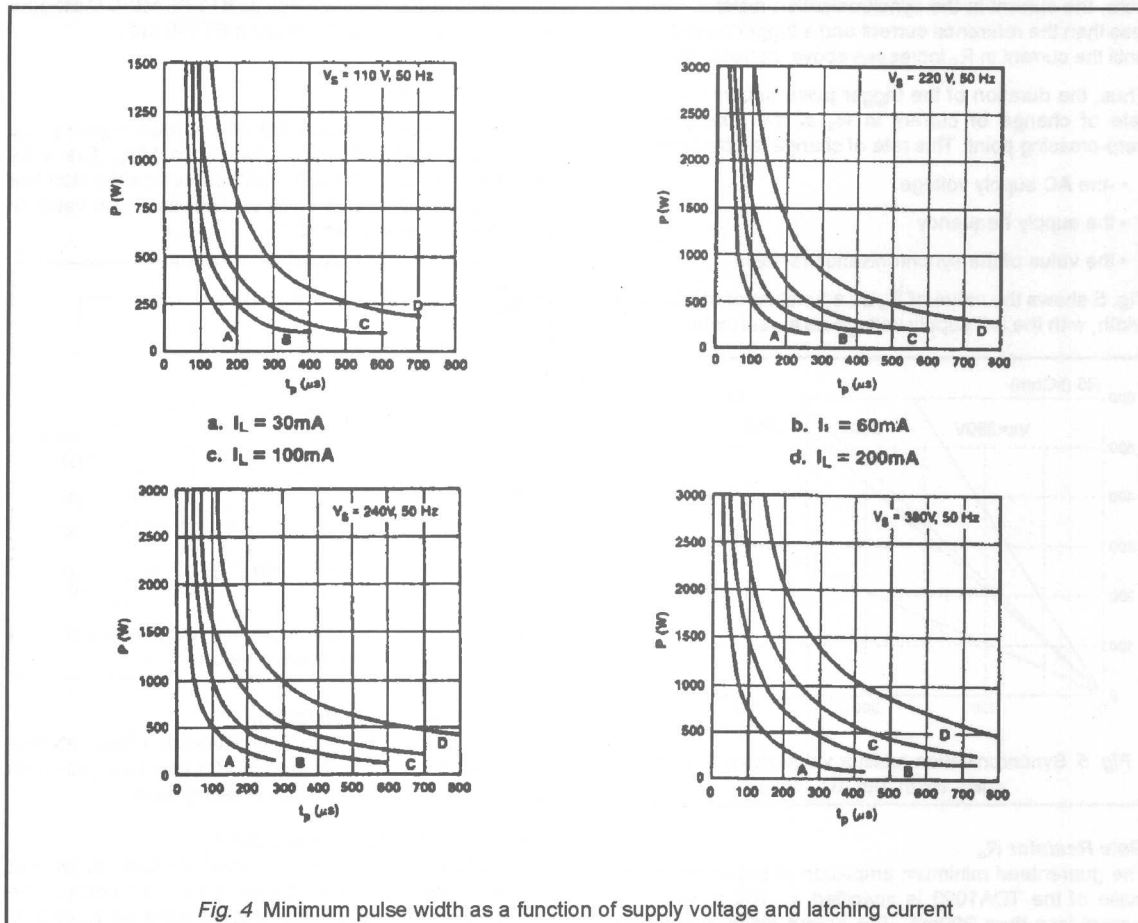


Fig. 3 TDA1023 block diagram and external components



In general, the latching current of a triac is higher than the holding current, so the minimum trigger pulse duration may be taken as twice the time for the load current in the triac ( $I_T$ ) to rise from zero to the triac's latching current. See Fig. 2. The current passed by the triac is a function of its on-state voltage, the load resistance, and the applied voltage. The trigger pulse width is therefore a function of:

- triac latching current ( $I_L$ )
- applied AC voltage ( $v = V_s \sin \omega t$ )
- load resistance ( $R$ )
- on-state voltage of the triac ( $V_T$ ) at  $I_L$ .

The load resistance is related to the nominal load power,  $P$  and nominal supply voltage,  $V_s$  by  $R = V_s^2 / P$ . Assuming that the load resistance has a tolerance of 5% and the AC voltage variation is 10%, the minimum required width of the

trigger pulse in the worst case can be calculated. The graphs of Fig. 4 show  $t_{p(\text{MIN})}$  as a function of  $P$  for four common mains voltages with values of 30mA, 60mA, 100mA, and 200mA for the triac latching current  $I_L$  and a maximum on-state voltage of  $V_T = 1.2\text{ V}$  at  $I_L$ .

### Selection of external components

The external components required by the TDA1023 determine the operation of the device. The following paragraphs describe the selection of these components to ensure reliable operation under worst-case conditions.

#### Synchronisation Resistor, $R_s$

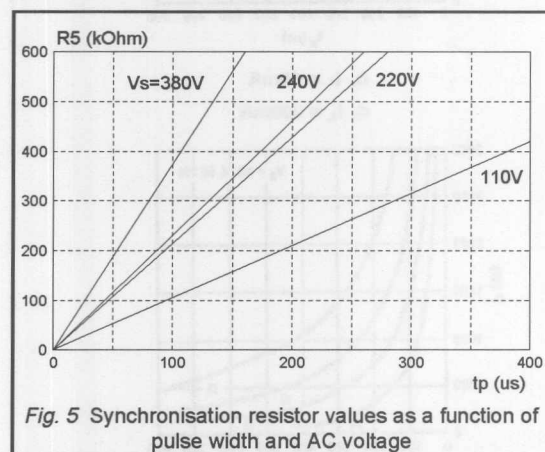
A current comparator is used as a zero-crossing detector to provide trigger pulse synchronisation. It compares the current through the synchronisation resistor ( $R_s$ ) with a reference current. As the supply voltage passes through

zero, the current in the synchronisation resistor becomes less than the reference current and a trigger pulse is given until the current in  $R_s$  increases above the reference level.

Thus, the duration of the trigger pulse depends upon the rate of change of current in  $R_s$  at the supply voltage zero-crossing point. This rate of change is affected by:

- the AC supply voltage
- the supply frequency
- the value of the synchronisation resistor.

Fig. 5 shows the value of  $R_s$  as a function of trigger pulse width, with the AC supply voltage as a parameter.



#### Gate Resistor $R_g$

The guaranteed minimum amplitude of the output trigger pulse of the TDA1023 is specified as 10V at an output current less than 200mA. The output stage is protected against damage due to short-circuits by current-limiting action when the current rises above 200mA.

Although the output is current-limited, it is still advantageous to include a gate series resistor in the circuit. Inclusion of a gate resistor to limit the gate current to the minimum value required reduces the overall current consumption and the power dissipation in the mains dropping resistor. Furthermore, the point at which current limiting occurs is subject to considerable variation between samples of the TDA1023: a gate resistor will reduce the effect of this in production circuits.

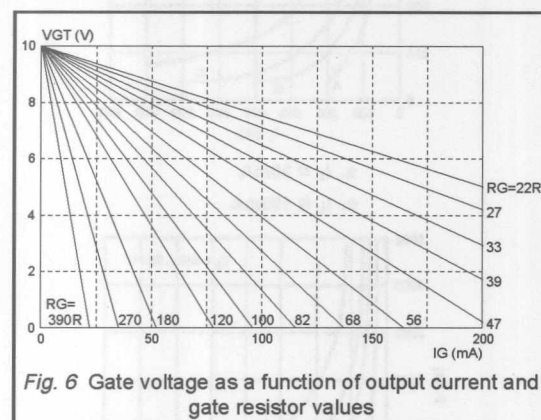
The rectangular output V/I characteristic of the TDA1023 is shown in Fig. 6. Load lines for various values of gate resistor have been plotted on this diagram so that the maximum value of gate resistor can be selected by plotting horizontal and vertical lines to represent the required minimum gate current and voltage. The following example illustrates the use of Fig. 6.

The triac to be triggered is a Philips BT139. At 0°C the trigger pulse requirements for a standard BT139 are:

$$I_{GT} = 98 \text{ mA}$$

$$V_{GT} = 1.6 \text{ V}$$

These figures are for triggering with a positive gate pulse when  $MT_2$  is negative with respect to  $MT_1$ . The lines representing  $V_{GT} = 1.6\text{V}$  and  $I_{GT} = 98\text{ mA}$  cross the load line for a gate resistor value of  $82\ \Omega$ . The maximum value of gate resistor is therefore  $82\ \Omega$ .



#### Gate Termination Resistor $R_{PD}$

The TDA1023 has a resistor approximately  $1.5\text{k}\Omega$  between Pin 1 and Pin 13. This is intended for use as a pull-down resistor when sensitive triacs are being used.

#### The Proportional Band Resistor $R_5$

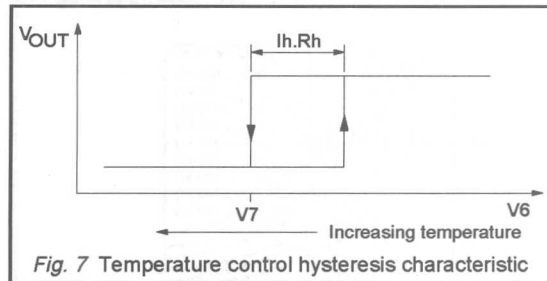
The proportional band is the input voltage range that provides control of 0% to 100% of the load power. The TDA1023 has a built-in proportional band of  $V_{pb} = 80\text{mV}$  (corresponding to about  $1^\circ\text{C}$ ) which can be increased by the addition of resistor  $R_5$  between Pin 5 and ground. The maximum proportional band of  $400\text{mV}$  is obtained by shorting Pin 5 to ground.

#### Hysteresis Resistor $R_4$

The comparator of the TDA1023 is designed with built-in hysteresis to eliminate instability and oscillation of the output which would cause spurious triggering of the triac. Apart from providing a stable two-state output, the hysteresis gives the comparator increased noise immunity and prevents half-waving.

Figure 7 shows the application of hysteresis to the comparator and the transfer characteristic obtained. The built-in hysteresis is  $20\text{mV}$ ; this may be increased by adding a resistor ( $R_4$ ) from Pin 4 to ground which increases the current  $I_H$ . Pin 4 shorted to ground gives a maximum of  $320\text{mV}$ . Table 2 gives the value of  $R_4$  for a range of hysteresis settings.





When the proportional band ( $V_{pb}$ ) is increased, it may be necessary to increase the hysteresis voltage ( $V_h$ ). Table 2 also shows a range of proportional band settings, the values of  $R_5$  required for these, the corresponding minimum hysteresis voltage and the maximum value of hysteresis resistor  $R_4$ .

Proportional Band (mV)	$R_5$ (k $\Omega$ )	$R_4$ (k $\Omega$ )	Hysteresis band (mV)
80	-	-	20
160	3.3	9.1	40
240	1.1	4.3	60
320	0.43	2.7	80
400	0.0	1.8	100

Table 2. Choice of components  $R_4$  and  $R_5$

Voltage (V)	AC rating ( $\mu$ F)	DC value ( $\mu$ F)	Catalogue Number
25	47	68	2222 016 90129
40	33	47	2222 016 90131
25	22	33	2222 015 90102
40	15	22	2222 015 90101
25	10	15	2222 015 90099
40	6.8	10	2222 015 90098

Table 3. Preferred capacitors for use with TDA1023

Power (W)	CENELEC $t_0$ (s)	$C_T$ (DC) ( $\mu$ F)	$t_{0(nom)}$ (s)	$t_{0(min)}$ (s)	$t_{0(max)}$ (s)
2000	24.0	68	41	22	65
1800	16.0	47	28	15	45
1600	10.0	33	20	11	32
1400	7.0	22	13	7	21
1200	4.6	15	9	4.8	14
1000	2.0	10	6	3.2	9.6
800	0.8	10	6	3.2	9.6
600	0.3	10	6	3.2	9.6

Table 4. Timing capacitor values for 220V operation

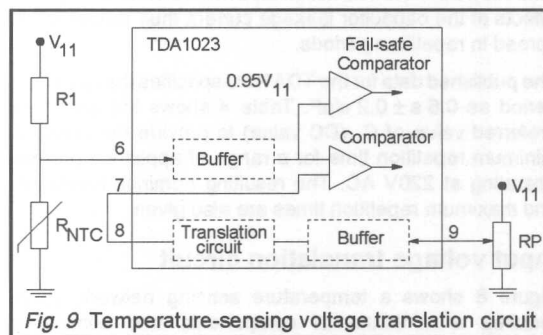
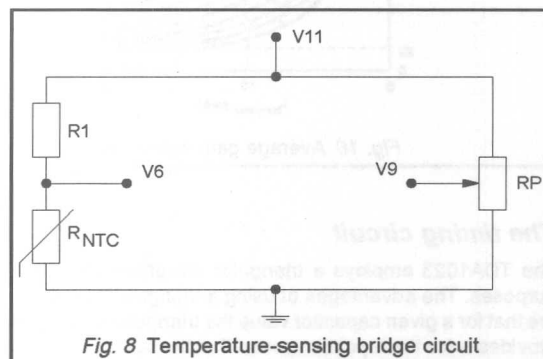
#### Smoothing Capacitor, $C_s$

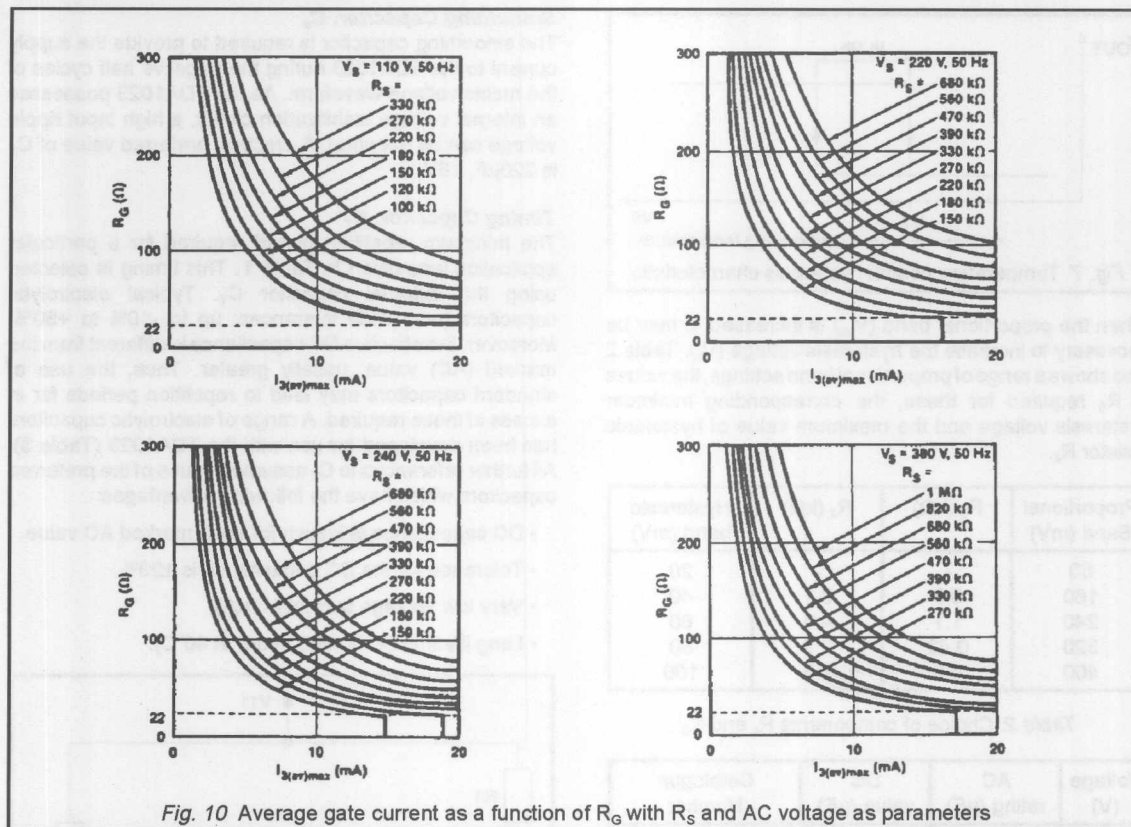
The smoothing capacitor is required to provide the supply current to the TDA1023 during the negative half cycles of the mains voltage waveform. As the TDA1023 possesses an internal voltage stabilization circuit, a high input ripple voltage can be tolerated. A practical preferred value of  $C_s$  is 220 $\mu$ F, 16V.

#### Timing Capacitor, $C_T$

The minimum repetition period required for a particular application was given in Table 1. This timing is selected using the external capacitor  $C_T$ . Typical electrolytic capacitors have wide tolerances: up to -10% to +50%. Moreover, the effective DC capacitance is different from the marked (AC) value, usually greater. Thus, the use of standard capacitors may lead to repetition periods far in excess of those required. A range of electrolytic capacitors has been developed for use with the TDA1023 (Table 3). All further references to  $C_T$  assume the use of the preferred capacitors which have the following advantages:

- DC capacitance is known for each marked AC value.
- Tolerance for the DC capacitance is  $\pm 20\%$ .
- Very low leakage current ( $< 1\mu$ A).
- Long lifetime ( $> 100,000$  hours at  $40^\circ\text{C}$ ).





### The timing circuit

The TDA1023 employs a triangular waveform for timing purposes. The advantages of using a triangular waveform are that for a given capacitor value the triangular waveform provides twice the repetition period that the sawtooth gives. This allows the use of smaller capacitors and minimises the effects of the capacitor leakage current thus reducing the spread in repetition periods.

The published data for the TDA1023 specifies the repetition period as  $0.6 \pm 0.2$  s/ $\mu$ F. Table 4 shows the minimum preferred value of  $C_T$  (DC value) to provide the required minimum repetition time for a range of appliance powers operating at 220V AC. The resulting nominal, minimum, and maximum repetition times are also given.

### Input voltage translation circuit

Figure 8 shows a temperature sensing network which requires a minimum of components and eliminates performance spreads due to potentiometer tolerances. For applications where the input voltage variation is very much

less than the available voltage then the required temperature will be controlled by a small angle of rotation of the potentiometer shaft. The TDA1023 voltage translation circuit allows the use of 80% of the potentiometer rotation giving accurate control of the temperature. If the voltage translation circuit is not used then pins 9 and 11 must be shorted together to disable the circuit. A block diagram of the translation circuit is shown in Fig. 9.

### Fail-safe circuits

The TDA1023 is fail-safe for both short-circuit and open-circuit conditions. Either of these conditions will prevent production of trigger pulses for the triac.

Short-circuit sensing is automatically obtained from the normal temperature sensing circuit. When the thermistor input voltage is zero, the triac will never be triggered because the potentiometer slider voltage will be higher. To sense the open-circuit thermistor condition, an extra comparator is used. This fail-safe comparator will inhibit output pulses if the thermistor input voltage rises above a

reference value (see Fig. 9).

### Determination of required supply current

Before any calculations concerning the required supply current can be made, the maximum average output current of the TDA1023 must be determined. The minimum supply current required is the sum of the following currents:

- the maximum average output current
- the current drawn by the temperature-sense circuit
- the current required by the integrated circuit.

For worst-case conditions, a 5% tolerance for  $R_S$  and  $R_G$  and a 10% variation of the mains is assumed. Figure 10 shows graphs of  $I_{G(AV)max}$  as a function of  $R_G$  and  $R_S$  for four 50Hz supply voltages. Below  $R_G=22\Omega$  there is no further increase in  $I_G$  as the output current is limited. The current drawn by the temperature-sensing circuit must not be greater than 1mA. The current consumption of the

TDA1023 depends upon the hysteresis and proportional band settings. Figure 11 shows the minimum supply current as a function of the average output current for limit settings.

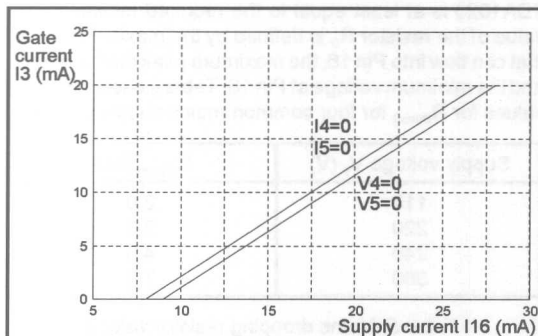


Fig. 11 Maximum required input current as function of gate current for limits of hysteresis band settings

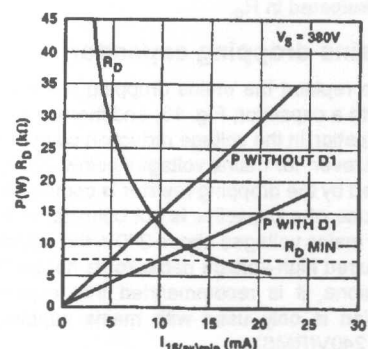
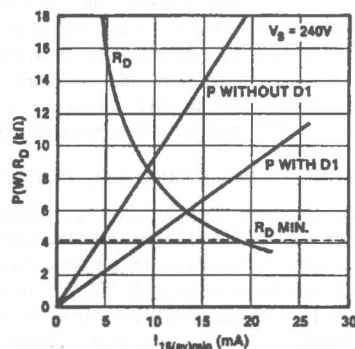
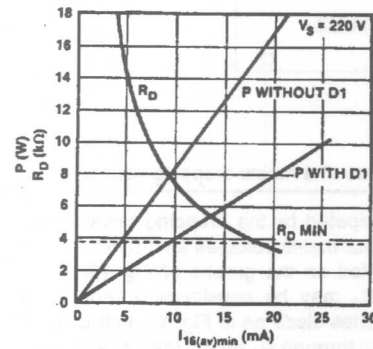
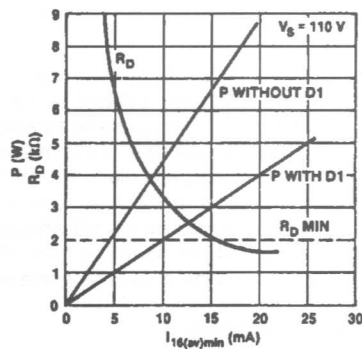


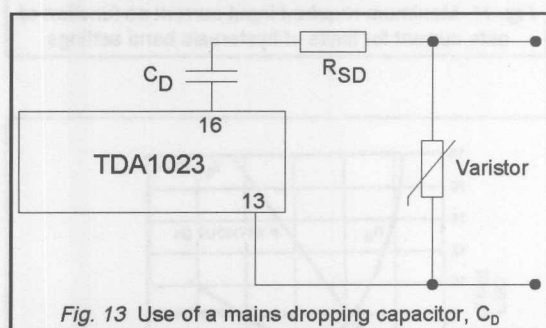
Fig. 12 Input current as a function of  $R_D$  and power dissipation (with and without series diode)

**The mains dropping resistor,  $R_D$** 

The value of the mains dropping resistor must be chosen such that the average supply current to the input of the TDA1023 is at least equal to the required minimum. The value of the resistor  $R_D$  is defined by the maximum current that can flow into Pin 16, the maximum peak mains voltage, and the minimum voltage at Pin 16. Table 5 shows practical values for  $R_{D(min)}$  for four common mains supply voltages

Supply voltage $V_s$ (V)	$R_{D(min)}$ (k $\Omega$ )
110	2.0
220	3.9
240	4.3
380	7.5

Table 5. Mains dropping resistor values

Fig. 13 Use of a mains dropping capacitor,  $C_D$ 

The power dissipated by the dropping resistor has been computed for four mains voltages as a function of  $R_D$  and the results plotted on the graphs of Fig. 12. The power dissipated in  $R_D$  may be considerably reduced by the addition of a series diode as in Fig. 14. In this case there is no conduction through  $R_D$  during the negative half-cycle of the supply voltage, giving a reduction of more than 50% of the power dissipated in  $R_D$ .

**Use of a mains dropping capacitor**

It is possible to replace the mains dropping resistor and series diode with a capacitor, Fig. 13, and thereby reduce the power dissipation in the voltage reduction components still further. However, for mains voltages below 200V, the power dissipated by the dropping resistor is comparatively small and the use of a capacitor is not considered to be necessary. For mains voltages above 240V, the additional cost of the required high-voltage capacitor is not justified. For these reasons, it is recommended that capacitive voltage reduction is only used with mains supplies of 200V(RMS) or 240V(RMS).

When selecting a capacitor for mains voltage reduction, the following points must be considered:

- **AC voltage rating**

- **Suppression of mains-borne transients** - A voltage-dependent resistor must be connected across the mains input to limit mains borne transients. For  $R_{SD}=390\Omega$  this yields a maximum transient voltage of about 740V. For 220V operation, a VDR (catalogue number 2322 594 13512) will limit the supply voltage to the required level during current transients of up to about 200A. For 240V operation, a VDR (catalogue number 2322 594 13912) will limit the supply voltage to the required level during current transients of up to about 80A.

- **Limit of inrush current** - The capacitor  $C_D$  must not be chosen so large that the input current to the TDA1023 violates the absolute maximum specified in the published data. A practical value for  $C_D$  is 680nF. Resistor  $R_{SD}$  must also limit the peak value of the inrush current to less than 2A under worst case operating conditions. With a 240V (+10%) supply, the value of  $390\Omega$  (-5%) will limit the worst case peak value of the inrush current to:

$$\frac{240 \times 1.1}{0.95 \times 390} \sqrt{2} = 1.01A$$

**Triac protection**

If the mains dropping circuit consists of capacitor  $C_D$  and resistor  $R_{SD}$ , a VDR must be included in the circuit as described above. This VDR will also protect the triac against current surges in the mains supply. If the mains dropping circuit consists of resistor  $R_D$  and diode D1, the VDR may be connected directly across the triac, giving improved protection due to the series resistance of the heater. Current surges in the supply will not harm the TDA1023 as the dropping resistor will limit the current to a safe level.

**Application examples**

The TDA1023 is intended primarily for room temperature control using electric panel heaters. The controllable heater power range is from 400W to 2000W, although the upper limit may be increased by suitable choice of triacs and/or heatsinks. The TDA1023 may also be used as a time proportional switch for cooker elements and similar devices, giving 100% control of the power dissipation.

**1. Domestic panel heater controller**

Figure 3 showed the design for a time proportional heater control using the TDA1023. Economies may be gained by the use of smaller or lower power components and so two versions are described in Table 6. Version A, for heaters from 400W to 1200W, uses a BT138 triac and a 15 $\mu$ F timing capacitor, version B, for heaters from 1200W to 2000W, uses a BT139 triac and a 68 $\mu$ F timing capacitor. Table 6 gives the necessary component values under worst case conditions for each of these versions for use with mains supplies of 220V, 50Hz.



The capacitor C1 has been included in the circuit of Fig. 3 to minimise sensor line interference pick-up. This is only necessary when the sensor is remote from the control circuit. The built-in hysteresis and proportional band provides optimum performance for panel heaters so pins 4 and 5 are not connected.

Component	Version A 400W - 1200W	Version B 1200W - 2000W
T <sub>1</sub>	BT138-500	BT139-500
VDR <sup>1</sup>	350V, 1mA	350V, 1mA
D <sub>1</sub>	BYX10G	BYX10G
R <sub>1</sub> <sup>2</sup>	18.7kΩ	18.7kΩ
R <sub>NTC</sub> <sup>3</sup>	R25=22kΩ, B=4200k	R25=22kΩ, B=4200k
R <sub>p</sub>	22kΩ	22kΩ
R <sub>D</sub>	4.3kΩ	6.2kΩ
R <sub>G</sub>	82Ω	82Ω
R <sub>S</sub>	430kΩ	180kΩ
C <sub>1</sub>	47nF	47nF
C <sub>S</sub>	220μF, 16V	220μF, 16V
C <sub>T</sub>	15μF (DC)	68μF (DC)
C <sub>D</sub> <sup>4</sup>	680nF	470nF
R <sub>SD</sub> <sup>4</sup>	390Ω	390Ω

Notes: 1. Cat. No. 2322 594 13512  
2. 1% tolerance  
3. Cat. No. 2322 642 12223  
4. Only required if used in place of D<sub>1</sub> and R<sub>D</sub>

Table 6. 220V, 50Hz temperature controller components

## 2. Temperature control of 2kW load.

For a load power of 2kW the BT139 triac must be used. The circuit is also that shown in Fig. 3. Table 7 gives a summary of the required component values.

Component	Value	Remarks
T <sub>1</sub>	BT139-500	
VDR	350V, 1mA	No. 2322 594 13912
D <sub>1</sub>	BYX10G	
R <sub>1</sub>	18.7kΩ	1% tolerance
R <sub>NTC</sub>	R25=22kΩ, B=4200k	No. 2322 642 12223
R <sub>p</sub>	22kΩ	
R <sub>D</sub>	6.8kΩ	
R <sub>G</sub>	82Ω	
R <sub>S</sub>	150kΩ	
C <sub>1</sub>	47nF	
C <sub>S</sub>	220μF, 16V	
C <sub>T</sub>	47μF (DC)	No. 2222 016 90129

Table 7. 2000W, 220V, 50Hz temperature controller

### Value of R<sub>S</sub>

The required trigger pulse width can be found from Fig. 4 as a function of the load power, latch current and supply voltage (2000W, 60mA, and 220V, 50Hz, respectively):  $t_{p(min)}=64\mu s$ . A value of  $R_S=135k\Omega$  provides a trigger pulse of the required duration. The next preferred value above this is 150kΩ, providing a  $t_{p(min)}$  of approximately 70μs.

### Value of R<sub>G</sub>

The maximum value of R<sub>G</sub> that may be used is determined by the minimum conditions to reliably trigger all samples of the triac. In Fig. 6 it can be seen that the operation point of 1.6V and 98 mA lies on the load line for 82Ω and this is the value chosen.

### Value of C<sub>T</sub>

For a load of 2kW, the repetition period must be at least 24s (from Table 1). From Table 4 the minimum preferred value of C<sub>T</sub> to provide this period is 68μF. However, due to the different performance under AC and DC conditions, then from Table 3, the actual capacitor used should be 47μF, 25V.

### Value of R<sub>1</sub> and R<sub>p</sub>

For control over the range 5°C to 35°C and a thermistor characteristics with  $R_{25}=22k\Omega$ , a suitable value of R<sub>1</sub> is 18.7kΩ ±1%. A suitable value for R<sub>p</sub> is 22kΩ.

### Value of R<sub>D</sub>

First, the maximum average output current must be found. From Fig. 10 the maximum gate current I<sub>G</sub> is given as a function of the values of resistors R<sub>S</sub> and R<sub>G</sub>. For this circuit  $I_{G(AV)max}=5$  mA. Once the maximum average output current is known, the minimum required supply current can be found from Fig. 11. With minimum hysteresis and proportional band, the average value of the supply current is 12.5 mA. Using this value of input current the required value of R<sub>D</sub> can be found from Fig. 12 giving R<sub>D</sub> = 5.6 kΩ. The power dissipation in the resistor when diode D<sub>1</sub> is present in the circuit is then 5 W.

## 3. Time proportional power control

The TDA1023 may be used to provide proportional control of devices such as electric cooker elements. The temperature-sensing bridge is replaced by a potentiometer, the power in the load being proportional to the potentiometer setting. Proportional power control is thus obtained while the potentiometer voltage lies between the upper and lower limits of the triangular waveform comparator input.

As the timing capacitor is charged and discharged by current sources, the voltage across it will never reach zero, so that load power will be zero before the potentiometer reaches its minimum setting. Similarly, maximum load power is reached before the maximum setting of the potentiometer. This effect can be reduced by the addition of resistors R<sub>1</sub> and R<sub>2</sub>. To ensure that 0% and 100% load

power can be selected by the potentiometer setting, the values of  $R_1$  and  $R_2$  should each be limited to 10% of the value of  $R_p$ .

All the circuit components are calculated in the same way as for the temperature controller, including the timing capacitor  $C_T$ . An example circuit, with components suitable for the control of loads from 1kW to 2kW from 220V, 50Hz supplies, is shown in Fig. 14 and Table 8.

Component	Value
$T_1$	BT139-500
VDR	ZnO, 350V, 1mA
$D_1$	BYX10G
$R_1$	4.7k $\Omega$
$R_2$	4.7k $\Omega$
$R_p$	47k $\Omega$
$R_D$	5.6k $\Omega$
$R_G$	82 $\Omega$
$R_S$	220k $\Omega$
$C_S$	220 $\mu$ F, 16V
$C_T$	47 $\mu$ F, 25V

Table 8. Time proportional power controller

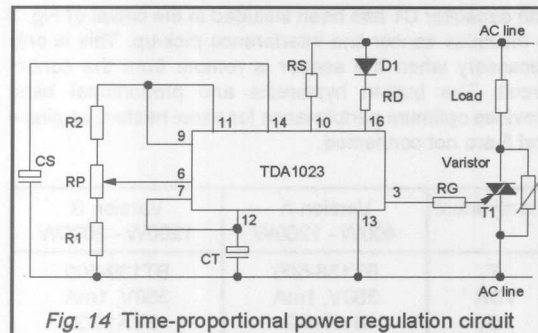
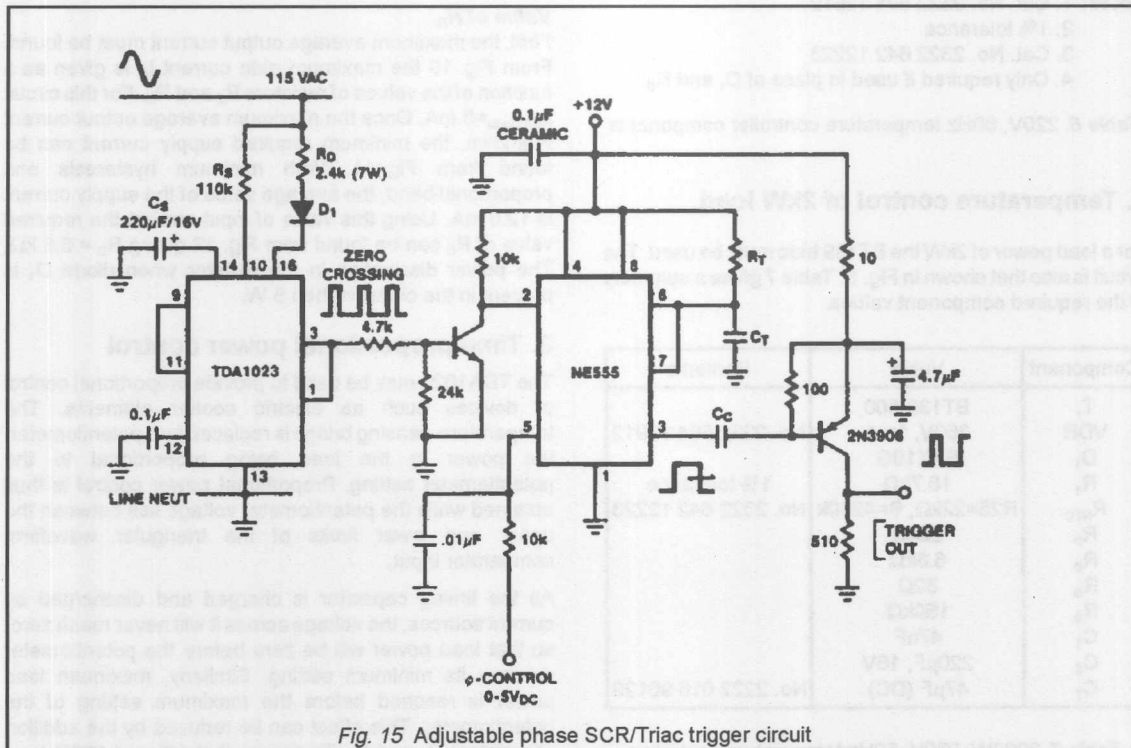


Fig. 14 Time-proportional power regulation circuit

#### 4. Phase control circuit using the TDA1023

Figure 15 shows an adjustable phase control trigger circuit suitable for thyristor or triac controller applications. The circuit uses the TDA1023 control chip and an NE555 timer device to give output phase control proportional to the input voltage command.



*Hi-Com Triacs*





### 6.3.1 Understanding Hi-Com Triacs

Hi-Com triacs from Philips Semiconductors are specifically designed to give superior triac commutation performance in the control of motors for domestic equipment and tools. These devices are suitable for use with a wide variety of motor and inductive loads without the need for a protective snubber. The use of a Hi-Com triac greatly simplifies circuit design and gives significant cost savings to the designer.

This product information sheet explains how the superior characteristics and performance of Hi-Com triacs removes design limitations of standard devices.

#### Triac commutation explained

A triac is an AC conduction device, and may be thought of as two antiparallel thyristors monolithically integrated onto the same silicon chip.

In phase control circuits the triac often has to be triggered into conduction part way into each half cycle. This means that at the end of each half cycle the on-state current in one direction must drop to zero and not resume in the other direction until the device is triggered again. This "commutation" turn-off capability is at the heart of triac power control applications.

If the triac were truly two separate thyristors this requirement would not present any problems. However, as the two are on the same piece of silicon there is the possibility that the "reverse recovery current" (due to unrecombined charge carriers) of one thyristor as it turns off, may act as gate current to trigger the other thyristor as the voltage rises in the opposite direction. This is described as a "commutation failure" and results in the triac continuing to conduct in the opposite direction instead of blocking.

The probability of any device failing commutation is dependent on the rate of rise of reverse voltage ( $dV/dt$ ) and the rate of decrease of conduction current ( $dI/dt$ ). The higher the  $dI/dt$  the more unrecombined charge carriers are left at the instant of turn-off. The higher the  $dV/dt$  the more probable it is that some of these carriers will act as gate current. Thus the commutation capability of any device is usually specified in terms of the turn-off  $dI/dt$  and the re-applied  $dV/dt$  it can withstand, at any particular junction temperature.

If a triac has to be operated in an inductive load circuit with a combination of  $dI/dt$  and  $dV/dt$  that exceeds its specification, it is necessary to use an RC-snubber network in parallel with the device to limit the  $dV/dt$ . This is at a penalty of extra circuit complexity and dissipation in the snubber. The "High Commutation" triacs (Hi-Com triacs) are designed to have superior commutation capability, so that even at a high rate of turn-off ( $dI/dt$ ) and a high rate of re-applied  $dV/dt$  they can be used without the aid of a

snubber network, thus greatly simplifying the circuit. The design features of Hi-Com devices that have made this possible are:

#### Geometric separation of the two antiparallel thyristors

Commutation failure can be avoided by physically separating the two 'thyristor halves' of a triac. However, separating them into two discrete chips would remove the advantage of a triac being triggerable in both directions by the same gate connection. Within the integrated structure of a Hi-Com triac the two halves of the device are kept further apart by modifying the layout of the chip in order to lessen the chance of conduction in one half affecting the other half.

#### Emitter shorting

"Emitter shorts" refer to the on-chip resistive paths between emitter and base of a transistor. A higher degree of emitter shorting means the presence of more such paths and lower resistance values in them. The use of emitter shorts in a triac has two effects on commutation.

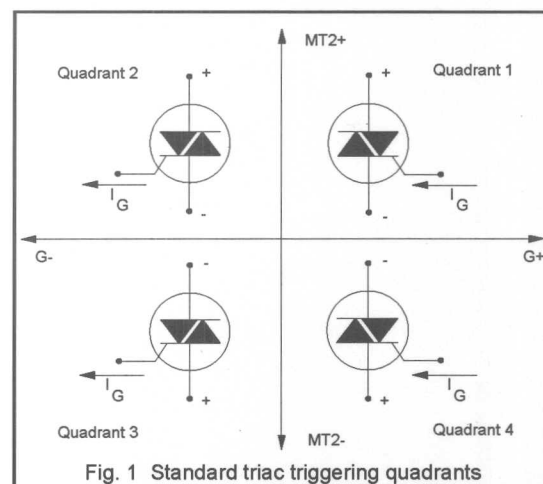


Fig. 1 Standard triac triggering quadrants

Firstly it reduces the gain of the internal transistors that make up the triac. This means there will be fewer carriers left to recombine when the conduction current falls to zero, and therefore a smaller probability that a sufficient number will be available to re-trigger the triac. The second way in which emitter shorts help commutation is that any unrecombined carriers in the conducting thyristor at turn-off

will have more chance of flowing out through the emitter shorts (of the opposite thyristor) rather than acting as gate current to trigger that thyristor on.

The Hi-Com triacs have a higher degree of emitter shorting both around the periphery of the device and in the central part of the active area. This both reduces the number of carriers available, and lessens the danger of any available carriers acting as gate current for undesirable triggering.

### Modified gate structure

The gate of a triac allows conduction in both directions to be initiated by either a positive or a negative current pulse between gate (G) and main terminal (MT1). The four different modes of triggering are often called 1+, 1-, 3- and 3+ (or sometimes quadrants 1, 2, 3 and 4) and are shown in Fig. 1.

This triggering versatility arises from the fact that the gate consists of some elements which conduct temporarily during the turn-on phase. In particular, one of the triggering modes, 3+ (or quadrant 4), relies on the main terminal 1 supplying electrons to trigger a thyristor element in the gate-MT1 boundary. Conduction then spreads to the main thyristor element from this boundary.

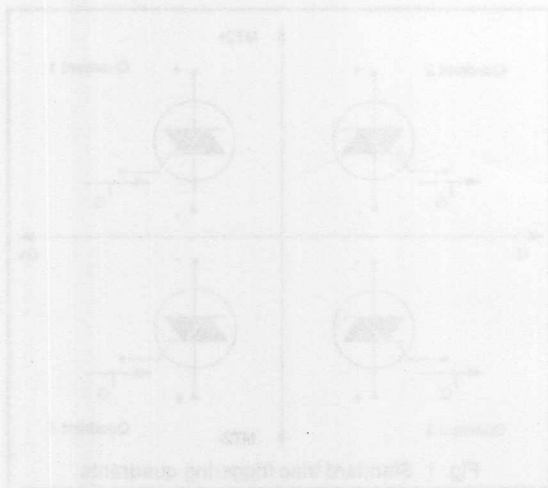
Unfortunately the carrier distribution in this triggering mode of operation is very similar to that existing when the triac is commutating in the 1-to-3 direction (i.e. changing from

conduction with MT2 positive to blocking with MT1 positive). The presence of the element in the gate to allow 3+ triggering will therefore always also undermine commutation capability in the 1-to-3 direction. For this reason the Hi-Com triacs have a modified gate design to remove this structure. This incurs the penalty that the 3+ trigger mode cannot be used, but it greatly improves the commutation performance of the device.

### Conclusions

By modifications to the internal design and layout of the triac it is possible to achieve a high commutation capability triac for use in inductive and motor load applications. These modifications have been implemented in the Hi-Com range of devices from Philips Semiconductors. The devices can be used in all typical motor control applications without the need for a snubber circuit. The commutation capability of the devices is well in excess of the operating conditions in typical applications.

As the loss of the fourth trigger quadrant can usually be tolerated in most designs, Hi-Com triacs can be used in existing motor control applications without the snubber network required for a standard device. This gives the designer significant savings in design simplicity, board space and system cost.



### 6.3.2 Using Hi-Com Triacs

Hi-Com triacs from Philips Semiconductors are specifically designed to give superior triac commutation performance in the control of motors for domestic equipment and tools. These devices are suitable for use with a wide variety of motor and inductive loads without the need for a snubber. The use of a Hi-Com triac greatly simplifies circuit design and gives significant cost savings to the designer.

This product information sheet explains how the need for a triac snubber arises and how the superior performance of Hi-Com triacs removes design limitations of standard devices. The Hi-Com range is summarised in Table 1.

#### Triac commutation

For resistive loads the device current is in phase with the line voltage. Under such conditions triac turn-off (commutation) occurs at the voltage "zero-crossover" point. This is not a very severe condition for triac commutation: the slow rising  $dV/dt$  gives time for the triac to turn off (commutate) easily.

The situation is quite different with inductive or motor loads. For these circuits conduction current lags behind the line voltage as shown in Fig. 1. When triac commutation occurs the rate of rise of voltage in the opposite direction can be very rapid and is governed by the circuit and device

characteristics. This high  $dV/dt$  means there is a much higher probability of charge carriers in the device re-triggering the triac and causing a commutation failure.

#### Hi-Com triacs

Hi-Com triacs are specifically designed for use with ac inductive loads such as motors. As commutation capability is not an issue for resistive load applications then standard triacs are still the most appropriate devices for these applications. The significant advantage of a Hi-Com triac is that it has no limitation on the rate of rise of reapplied voltage at commutation. This removes the requirement for a snubber circuit in inductive load circuits. An additional advantage of the Hi-Com design is that the off-state (static)  $dv/dt$  capability of the device is also significantly improved.

When using Hi-Com triacs in inductive load applications the trigger circuit cannot trigger the device in the fourth (3+) quadrant (Fig. 2). Fortunately the vast majority of circuit designs do not require this mode of operation and so are suitable for use with Hi-Com triacs without modification. The circuit of Fig. 3 is a typical example of the simplest type of trigger circuit. Hi-Com triacs are equally suitable for use with microcontroller trigger circuits.

Parameter		BTA212-600B	BTA212-800B	BTA216-600B	BTA216-800B
Repetitive peak voltage	$V_{DRM}$ (V)	600	800	600	800
RMS on-state current	$I_{T(RMS)}$ (A)	12	12	16	16
Gate trigger current	$I_{GT}$ (mA)	2 - 50	2 - 50	2 - 50	2 - 50
Off state $dv/dt$	$dV_D/dt$ (V/ $\mu$ s)	1000	1000	1000	1000
Commutating $di/dt$	$dI_{com}/dt$ (A/ms)	24	24	28	28
Turn-on $di/dt$	$dI_T/dt$ (A/ $\mu$ s)	50	50	50	50
Package		TO220	TO220	TO220	TO220

Table 1. Philips Semiconductors Hi-Com Triac range

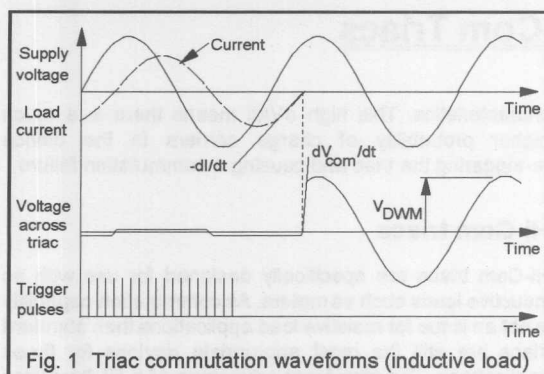


Fig. 1 Triac commutation waveforms (inductive load)

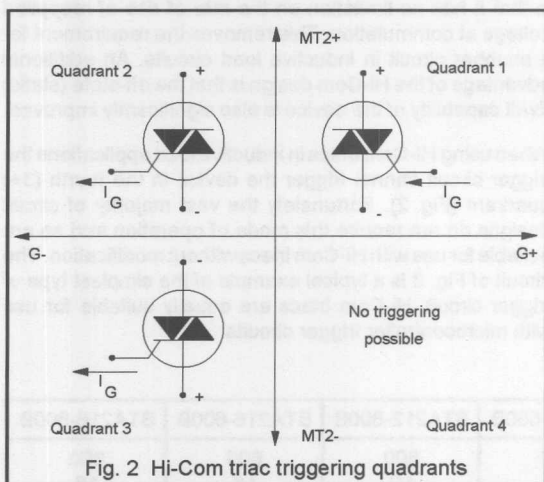


Fig. 2 Hi-Com triac triggering quadrants

### Device limiting values

#### i) Trigger current, $I_{GT}$

Trigger current for the Hi-Com triacs is in the range 2mA to 50mA. This means that gate currents due to noise that are below 2mA in amplitude can be guaranteed not to trigger the devices. This gives the devices a noise immunity feature that is important in many applications. The trigger current delivered by the trigger circuit must be greater than 50mA under all conditions in order to guarantee triggering of the device when required. As discussed above, triggering is

only possible in the 1+, 1- and 3- quadrants.

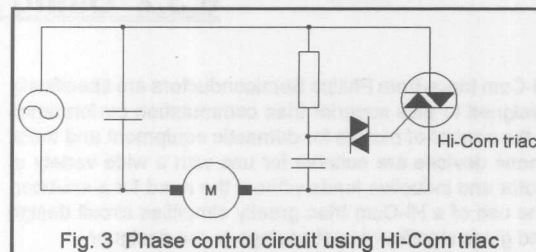


Fig. 3 Phase control circuit using Hi-Com triac

#### ii) Rate of change of current, $di_{com}/dt$

Hi-Com triacs do not require a snubber network providing that the rate of change of current prior to commutation is less than the rating specified in the device data sheet. This  $di_{com}/dt$  limit is well in excess of the currents that occur in the device under normal operating conditions, during transients such as start-up and faults such as the stalled motor condition.

For the 12A Hi-Com triacs the limit commutating current is typically 24A/ms at 125°C. This corresponds to an RMS current of 54A at 50Hz. For the 16A Hi-Com triacs the limit commutating current is typically 28A/ms at 125°C. This corresponds to an RMS current of 63A at 50Hz. Typical stall currents for an 800W domestic appliance motor are in the range 15A to 20A and so the commutation capability of the Hi-Com triacs is well above the requirement for this type of application.

### Conclusions

The Hi-Com range of devices from Philips Semiconductors can be used in all typical motor control applications without the need for a snubber circuit. The commutation capability of the devices is well in excess of the operating conditions in typical applications.

As the loss of the fourth trigger quadrant can usually be tolerated in most designs, Hi-Com triacs can be used in existing motor control applications. By removing the snubber the use of a Hi-Com triac gives the designer significant savings in design simplicity, board space and system cost.



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## Preface

This book was prepared by the Power Semiconductor Applications Laboratory of the Philips Semiconductors product division, Hazel Grove. The book is intended as a guide to using power semiconductors both efficiently and reliably in power conversion applications. It is made up of eight main chapters each of which contains a number of application notes aimed at making it easier to select and use power semiconductors.

**CHAPTER 1** forms an introduction to power semiconductors concentrating particularly on the two major power transistor technologies, Power MOSFETs and High Voltage Bipolar Transistors.

**CHAPTER 2** is devoted to Switched Mode Power Supplies. It begins with a basic description of the most commonly used topologies and discusses the major issues surrounding the use of power semiconductors including rectifiers. Specific design examples are given as well as a look at designing the magnetic components. The end of this chapter describes resonant power supply technology.

**CHAPTER 3** describes motion control in terms of ac, dc and stepper motor operation and control. This chapter looks only at transistor controls, phase control using thyristors and triacs is discussed separately in chapter 6.

**CHAPTER 4** looks at television and monitor applications. A description of the operation of horizontal deflection circuits is given followed by transistor selection guides for both deflection and power supply applications. Deflection and power supply circuit examples are also given based on circuits designed by the Product Concept and Application Laboratories (Eindhoven).

**CHAPTER 5** concentrates on automotive electronics looking in detail at the requirements for the electronic switches taking into consideration the harsh environment in which they must operate.

**CHAPTER 6** reviews thyristor and triac applications from the basics of device technology and operation to the simple design rules which should be followed to achieve maximum reliability. Specific examples are given in this chapter for a number of the common applications.

**CHAPTER 7** looks at the thermal considerations for power semiconductors in terms of power dissipation and junction temperature limits. Part of this chapter is devoted to worked examples showing how junction temperatures can be calculated to ensure the limits are not exceeded. Heatsink requirements and designs are also discussed in the second half of this chapter.

**CHAPTER 8** is an introduction to the use of high voltage bipolar transistors in electronic lighting ballasts. Many of the possible topologies are described.

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